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Using Pulse Width Modulation for Wireless Transmission of Neural Signals in Multichannel Neural Recording Systems

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Abstract

We have used a well-known technique in wireless communication, pulse width modulation (PWM) of time division multiplexed (TDM) signals, within the architecture of a novel wireless integrated neural recording (WIneR) system. We have evaluated the performance of the PWM-based architecture and indicated its accuracy and potential sources of error through detailed theoretical analysis, simulations, and measurements on a setup consisting of a 15-channel WIneR prototype as the transmitter and two types of receivers; an Agilent 89600 vector signal analyzer and a custom wideband receiver, with 36 and 75 MHz of maximum bandwidth, respectively. Furthermore, we present simulation results from a realistic MATLAB-Simulink model of the entire WIneR system to observe the system behavior in response to changes in various parameters. We have concluded that the 15-ch WIneR prototype, which is fabricated in a 0.5- μ m standard CMOS process and consumes 4.5 mW from ± 1.5 V supplies, can acquire and wirelessly transmit up to 320 k-samples/s to a 75-MHz receiver with 8.4 bits of resolution, which is equivalent to a wireless data rate of ~ 2.26 Mb/s.

Keywords

Frequency shift keying; implantable microelectronic devices; neural interfacing; pulse width modulation; telemetry; time division multiplexing

I. Introduction

RAPID advancements in electrophysiology and behavioral neuroscience have resulted in a better understanding of the underlying principles that may describe the functional mechanisms of the human brain and some of the root causes of malfunctioning in its neuronal circuits [1]. The results may even inspire other areas of science and technology such as computing [2]. Due to ethical, logistical, and economical reasons, the majority of the ongoing neuroscience research is conducted on animal models such as rats, guinea pigs, cats, and nonhuman primates. The accelerating pace of research has created a considerable demand for data acquisition systems that are capable of simultaneously recording neural signals from a large number of electrodes placed in the neural tissue of awake behaving animals [3]. These applications should, therefore, be considered in design and development of the new neural recording instrumentation, which may not necessarily be used in humans any time soon.

For decades, researchers have been using arrays of discrete low noise amplifiers and filters that are implemented in large racks along with analog to digital conversion (ADC), digital

signal processing (DSP), and visualization hardware and software, which are now commercially available [4], [5]. These amplifiers are connected to the electrodes placed in the animal brain through a bundle of thin and flexible wires, often through a preamplifier on the animal headstage.

Hardwired interconnects are simple and provide a wide bandwidth. The wires, however, can potentially affect the animal behavior by causing psychophysical tethering effects besides adding noise and motion artifacts. To facilitate animal mobility and eliminate tangling and twisting of the wires, often a motorized commutator is used on top of the enclosure between the headstage and other instruments. Commutator is a delicate mechanical component and one of the most expensive items in the system [4], [5]. It is also the bottleneck for achieving large channel counts in hardwired setups, and limit experiments to only one animal at a time due to twisting and tangling of the wires. In general, providing a natural and enriched environment for animal subjects in a hardwired setup is not feasible, either with or without a commutator [6].

Considering the above issues, wireless neural recording (WNR) systems would be highly desired in electrophysiology setups, provided that they can match or surpass their hardwired counterparts. Such systems consist of at least a transmitter and a receiver unit before DSP. The transmitter should be implanted inside or carried on the animal body. It is also responsible for conditioning (mainly amplification and filtering) of the acquired neural signals, and should include a power source with enough energy storage for the minimum duration of uninterrupted experiments. The receiver unit needs to have at least one antenna inside the animal enclosure particularly if it is shielded in a Faraday cage.

Size, power consumption, robustness, input referred noise, and bandwidth are the main concerns in developing WNR systems, and several research groups have tried to tackle them in different ways [7]–[15]. Nevertheless, WNR systems are still generally absent in electrophysiology labs. The majority of the neuroscientists that the authors have talked to are willing to adopt WNR systems if they can seamlessly substitute their current hardwired systems at a reasonable cost, setting up effort, maintenance, and additional support.

Judy *et al.* have employed commercial off-the-shelf components in their WNR system, particularly to establish the wireless link using the ZigBee standard [7]. Even though this method can significantly reduce the development time, and has the added benefit of complying with the Federal Communications Commission (FCC) regulations, the size and power overhead in general purpose components may lose their competitive edge in high channel counts.

Several researchers have tried to tackle the bandwidth limitation by processing the neural signals on the transmitter unit by extracting their key features, and only sending a compressed version of the neural data across the wireless link [8]–[11]. The most important piece of information in a neural signal is the timing of the spike events [16]. Hence, once spikes generated by a specific neuron are detected, one can just send their timing information as opposed to the entire waveform. The challenge, however, is that the recorded signal from each extra cellular recording site contains spikes from a handful of nearby neurons that are randomly dispersed around the site as well as those that are far away and their activities only contribute to background noise. Spike neural activities should, therefore, be identified from noise and carefully sorted based on their waveforms before they can be converted to “single-unit” activities. To make things even more complicated, there are also gradual changes in the waveforms of the same neurons over time.

The processing power needed for the state-of-the-art spike sorting algorithms that operate on multiple channels in real time has required neuroscientists to employ high performance

multi-core computing clusters [17]. Embedding a comparable amount of computational power and programmability on the transmitter unit does not seem to be feasible in near future. Thus, such architectures are only suitable for neuroprostheses applications, where simple and low power routines would be sufficient [18], [19]. Moreover, many neuroscientists are interested not only in the single-unit activities but also in low frequency components of the neural signals, known as local field potentials (LFP), which are representative of the collective activities of thousands of neurons [10], [20].

Harris *et al.* have tried to encode the neural signal amplitude above the noise level in a series of sharp pulses, which frequency is proportional to the signal amplitude. This is a low power encoding scheme and works well for a single or small number of channels. However, it is not clear how the pulses generated from different channels would be combined to be transmitted across the wireless link [12].

Mohseni *et al.* and Morizio *et al.* have combined the sampled neural signals from different channels using time division multiplexing (TDM) before wireless transmission. The advantage of this method is its simplicity and low power consumption. However, analog samples are susceptible to noise, and the transitions from one channel to another in short sampling periods can result in significant crosstalk among adjacent channels on the receiver side [13], [14].

To get a sense of how much information needs to be transferred across the wireless link, we should note that the neural signal spectrum spans from 0.1 Hz to 10 kHz. Hence, the Nyquist rate requires 20 kS/s per channel. Considering that recorded neural signals are often between 50 μ V to 1 mV, supply range of ± 1.5 V, and the fact that even in a high quality recording there is often $> 10 \mu$ V of background noise, a resolution of 8–10 bits should be sufficient in this application [9]. Therefore, at least 160 kb/s of bandwidth is needed for raw data per recording channel.

To overcome the aforementioned problems, we have adopted an architecture based on pulse width modulation (PWM) of TDM signals in development of a prototype 15-ch Wireless Integrated Neural Recording (WIneR) system [21]. This method combines simplicity, small size, and low power consumption of the analog systems with robustness against noise and interference of the digital systems, while preserving and transferring the entire neural signal information from all recording channels to the external unit. Section II describes the architecture of the WIneR system. Section III covers the theory of the wireless PWM-TDM technique, followed by simulation and measurement results in Section IV.

II. WIneR System Architecture

A. Implantable Transmitter Unit

Block diagram and die micrograph of the 15-ch prototype WIneR system are shown in Fig. 1 [15]. The extracellular electrical activities are amplified at a gain of 100 and filtered from 0.1 Hz to 10 kHz using an array of low-noise amplifiers (LNA) with capacitive feedback [22]. LNAs are also responsible for rejecting the large dc baseline created at the electrode-tissue interface. A 16:1 TDM combines 15 channels of amplified neural signals plus a constant reference voltage (Mark) at 320 kHz. Mark identifies the beginning of each frame of 16 samples and serves as both sync and calibration signals on the receiver end.

A sample and hold (S/H) circuit follows the TDM to stabilize samples for PWM. The PWM block compares the S/H output with a triangular waveform generator (TWG) output through a high speed rail-to-rail comparator, resulting in a PWM-TDM signal, which has only high and low levels similar to digital signals. At this stage, the neural signal amplitude

information is encoded in the PWM-TDM duty cycle, D , which is more robust against noise and interference, in a process known as analog-to-time conversion (ATC). Complexity and power consumption of a single comparator is far less than a high speed ADC that among other circuits requires a few comparators. The PWM-TDM signal then drives a voltage controlled oscillator (VCO) to generate a frequency shift keyed (FSK) carrier to be transmitted by a miniature antenna [23]. The power-on reset (POR) block resets the chip when the supply is below a certain voltage.

B. External Receiver Unit

The external unit includes a custom designed wideband FM receiver with 75 MHz bandwidth [24]. The receiver can be tuned to RF carriers from 50 MHz to 1 GHz. A mixer down converts the FSK carrier, received from the implantable unit, to intermediate frequencies (IF). The amplified IF-PWM-FSK signal can be directly digitized for demodulation steps to proceed in the digital domain in a DSP board or PC. However, the volume of data for high channel counts could be overwhelming. Alternatively, the IF-PWM-FSK signal is fed into an FM demodulator (eg. a discriminator) followed by an envelope detector, such that the received signal is converted back to the PWM-TDM waveform. This signal is turned into a series of digitized samples by measuring its pulse widths using a time-to-digital converter (TDC), which can be a high speed timer/counter [24]. The stream of digitized samples is queued in a buffer and transferred to the PC through a USB link. Finally, the original neural signals are demultiplexed back to 15 individual waveforms by knowing their relative positions with respect to the synchronizing “Mark” sample.

III. Evaluation of the Wireless PWM Technique

In this section, we evaluate the wireless PWM technique for WNR application by exploring its potential sources of noise and inaccuracy. Our goal is to indicating the signal-to-noise ratio (SNR) on the receiver side as well as the overall WInER system resolution. Various sources of error can be divided into those that are related to 1) the implantable transmitter unit and 2) the external receiver unit. Transmitter errors include errors in generating the triangular waveform, PWM comparator noise, offset, and hysteresis, and the VCO phase noise. We do not consider the noise and nonlinearity of the front-end LNA here, because they are not specific to the PWM technique. On the receiver side, error is mostly due to the bandwidth and internal noise. One should, however, note that the purpose of the receiver in this architecture is not to reconstruct the exact transmitted PWM waveform but to accurately measure the time intervals between every two successive transitions in the received FSK carrier frequency.

A. Implantable Transmitter Errors

1) PWM Noise—The PWM noise includes both the TWG noise and the PWM comparator noise. Fig. 3(a) shows a simplified schematic diagram of the PWM circuit. A complementary current source/sink (CCSS) linearly charges/discharges a capacitor C between V_{high} and V_{low} [25]. The resulting triangular wave is compared with V_{in} and generates the PWM pulse width

$$w = C V_{\text{in}} \left(\frac{1}{I_{\text{Source}}} + \frac{1}{I_{\text{Sink}}} \right). \quad (1)$$

The pulse width jitter can be expressed as

$$\frac{dw}{T} = D \left(\frac{dC}{C} - \frac{I_{\text{Sink}} I_{\text{Source}}}{I_{\text{Source}} + I_{\text{Sink}}} \left(\frac{dI_{\text{Source}}}{I_{\text{Source}}^2} + \frac{dI_{\text{Sink}}}{I_{\text{Sink}}^2} \right) \right) + \frac{dV_{\text{in}}}{V_{\text{high}} - V_{\text{low}}} \quad (2)$$

where T is the sampling period and $D = w/T$ is the PWM duty cycle. Variations in the TWG capacitor are often very small. dV_{in} is the equivalent input referred noise of the comparator. Since the noise on V_{high} and V_{low} will also affect dw , all comparator noises need to be considered. dI_{Source} and dI_{Sink} are the noise contributions of the CCSS. Although the finite output impedance of CCSS also contributes to dw , it mostly causes distortion, which can be compensated in calibration.

To calculate the power of the jitter, we assume comparators are identical with an input referred noise of

$$\frac{w_n^2 PWM}{T^2} = \left(\frac{DI_{\text{Sink}} I_{\text{Source}}}{I_{\text{Source}} + I_{\text{Sink}}} \right)^2 \left(\frac{i_{n,\text{Source}}^2}{I_{\text{Source}}^4} + \frac{i_{n,\text{Sink}}^2}{I_{\text{Sink}}^4} \right) + 3 \left(\frac{V_{n,\text{comp}}}{V_{\text{high}} - V_{\text{low}}} \right)^2 \quad (3)$$

where $i_{n,\text{Source}}^2$ and $i_{n,\text{Sink}}^2$ are the current noise for the CCSS current source and sink, respectively.

2) VCO Noise—VCO is the only block that should be driven by the PWM comparator to create a FSK-PWM-TDM carrier signal (see Fig. 1). In this process, the PWM spectrum is shifted from baseband to FSK frequencies, f_1 and f_2 , as shown in Fig. 4.

A single square pulse with width w can be described in time and frequency domains as [Fig. 4(a)]

$$\Pi\left(\frac{t}{w}\right) = \begin{cases} 0, & |t| > \frac{w}{2} \\ 1, & |t| < \frac{w}{2} \\ \frac{1}{2}, & |t| = \frac{w}{2} \end{cases} \quad (4)$$

$$F\left[\Pi\left(\frac{t}{w}\right)\right] = \frac{\sin\pi f w}{\pi f}.$$

For a simple analysis, let us assume V_{in} is constant and the PWM period is T . The analysis for more complicated PWM signals with variable pulse width can be found in [26]. The PWM pulse train can be written as [Fig. 4(b)]

$$\begin{aligned} \text{PWM}(t) &= \sum_{n=-\infty}^{+\infty} \Pi\left(\frac{t-nT}{w}\right) \\ \text{PWM}(f) &= \sum_{n=-\infty}^{+\infty} \frac{\sin\pi \frac{nw}{T}}{\pi \frac{n}{T}} \delta\left(f - \frac{n}{T}\right) \end{aligned} \quad (5)$$

where $\delta(f)$ is the delta function. The FSK-PWM spectrum can then be found by shifting (5) $\pm f_1$ to $\pm f_2$ and [Fig. 4(c)]

$$\begin{aligned} \text{FSK-PWM}(t) &= \sum_{n=-\infty}^{+\infty} \Pi\left(\frac{t-nT}{w}\right) A \cos 2\pi f_1 t + \left(1 - \Pi\left(\frac{t-nT}{w}\right)\right) A \cos 2\pi f_2 t \\ \text{FSK-PWM}(f) &= \frac{A}{2} \sum_{n=-\infty}^{+\infty} \delta\left(f \pm f_1 - \frac{n}{T}\right) \frac{\sin\pi \frac{nw}{T}}{\pi \frac{n}{T}} - \frac{A}{2} \sum_{n=-\infty}^{+\infty} \delta\left(f \pm f_2 - \frac{n}{T}\right) \frac{\sin\pi \frac{nw}{T}}{\pi \frac{n}{T}} + \frac{A}{2} \delta(f \pm f_2). \end{aligned} \quad (6)$$

In the frequency domain, the VCO output will be the convolution of (6) and the VCO phase noise spectrum, $S_{\phi}(f)$, at an offset frequency, f , away from the carrier, f_{osc} . After receiving, down converting, and filtering of the FSK-PWM-TDM, if we ignore the bandwidth

limitation for the time being, the restored signal in the time domain will be the ideal PWM pulses in (5), shaped by the phase noise function

$$\text{PWM}_{Rx}(t) = \sum_{n=-\infty}^{+\infty} \prod \left(\frac{t-nT}{W} \right) F^{-1} [S_{\Phi}(f)]. \quad (7)$$

For white noise sources, frequency stability of the VCO is often characterized by the relative jitter [27],

$$S_{\Phi}(f) \cong \frac{f_{\text{osc}}}{f^2} \left(\frac{\Delta\tau_{\text{rms}}}{T_{\text{osc}}} \right)^2 \quad (8)$$

where $\Delta\tau_{\text{rms}}$ is the rms jitter, f_{osc} and T_{osc} are VCO carrier frequency and period. Using (8), it is easy to prove that

$$\left(\frac{\Delta f_{\text{rms}}}{f_{\text{osc}}} \right)^2 = \left(\frac{\Delta\tau_{\text{rms}}}{T_{\text{osc}}} \right)^2 = S_{\Phi}(f) \frac{f^2}{f_{\text{osc}}}. \quad (9)$$

The VCO phase noise results in a frequency noise of Δf_{rms} , which after frequency demodulation turns into an rms voltage noise, ΔV_{rms} . Due to the rising and falling slopes of the recovered PWM signal, ΔV_{rms} causes a pulse width error of ΔT_{VCO} , which can be found from

$$\frac{\Delta T_{\text{VCO}}}{T_{f/r}} = \frac{\Delta V_{\text{rms}}}{V_{pp}} = \frac{\Delta f_{\text{rms}}}{f_2 - f_1} \quad (10)$$

where $T_{f/r}$ is the sum of the rise and V_{pp} fall times and is the peak to peak voltage of the recovered PWM signal. Using (9) and (10), the PWM duty cycle error will be

$$\Delta D = \frac{\Delta T_{\text{VCO}}}{T} = \frac{f T_{f/r} \sqrt{S_{\Phi}(f) f_{\text{osc}}}}{T (f_2 - f_1)} \quad (11)$$

where f is the offset frequency at which VCO phase noise has been measured. Therefore, increasing the FSK modulation index and reducing $T_{f/r}$ can help reducing the VCO error.

Gradual VCO drift does not affect the FSK-PWM signal because it does not affect the pulse width. VCO settling time, however, is an important parameter determined by the VCO bandwidth and phase margin, which in turn depend on the VCO's tail current, LC tank quality factor, and loading.

B. External Receiver Errors

1) Receiver Thermal Noise—Maximum noise power transfer happens when there is impedance matching between successive blocks. This is usually the case in commercial devices since they are mostly designed for 50Ω. Matching between the receiver antenna and front-end RF LNA results in transferred noise power of

$$P_n = kT\Delta f \quad (12)$$

where Δf is the receiver RF front-end bandwidth. At room temperature, (12) in dBm becomes

$$P_{n,dBm} = [-174 + 10\log(\Delta f)] \text{ dBm}. \quad (13)$$

Every stage in Fig. 2 has a thermal noise characterized by its noise factor, F , and noise figure $NF = 10\log(F)$. If several stages with the gains of $G_1, G_2, \dots, G_n$ and noise factors of $F_1, F_2, \dots, F_n$ are connected in series, then

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots + \frac{F_n - 1}{G_1 G_2 \dots G_n}. \quad (14)$$

If the first two stages have a high enough gain, considering the first 3 terms in (14) would be sufficient.

Thermal noise for a receiver with 50Ω input impedance that is connected to an antenna with 50Ω radiation resistance would be

$$\begin{aligned} P_{n,Rx,dBm} &= P_{n,dBm} + 10\log(F - 1) \\ &= [-174 + 10\log(\Delta f) + 10\log(F - 1)] \text{ dBm}. \end{aligned} \quad (15)$$

If the input signal RF power (P_{sig}) is known, we can find the signal-to-noise ratio (SNR) for each stage. For PWM signal, the SNR is the reciprocal of pulse width duty cycle error.

$$SNR_{PWM} = \left(\frac{T}{\Delta W} \right)^2 = \frac{1}{\Delta D^2}. \quad (16)$$

Hence, the total receiver thermal noise contribution to the pulse width duty cycle error would be as shown in (17) at the bottom of the page.

2) Local Oscillator Phase Noise—The amplified FSK-PWM signal is down converted to IF- FSK-PWM by multiplication with a local oscillator (f_{LO}). The resulting signal will be similar to (6) when shifting the carrier frequencies from f_1 and f_2 to $f_1 - f_{LO}$ and $f_2 - f_{LO}$ [Fig. 5(a)], and the error analysis will be similar to the VCO in Section III-A-II. However in this application, the commercial LO often has a much lower phase noise compared to the transmitter VCO, and the LO contribution to pulse width jitter can be ignored.

3) Receiver Bandwidth (RBW) Limitation—Edges of the recovered PWM signal become rounded due to RBW limitation, introducing pulse width error to the system. We simplify our analysis by considering the receiver as an ideal low pass filter with cutoff frequency of f_{BW} , which limits the PWM spectrum to $-f_{BW} \sim f_{BW}$ range. Fig. 5 shows the received IF and baseband PWM signal in time and frequency domains. Mathematically, the complete PWM spectrum in (5) is multiplied by the receiver low pass filter

$$\begin{aligned} \Delta D &= \sqrt{\frac{1}{SNR_{PWM}}} = 10^{(P_{n,Rx,dBm} - P_{sig,Rx,dBm})/20} \\ &\approx 10^{\left[-174 + 10\log(\Delta f) + 10\log(F_1 + (F_2 - 1/G_1) + (F_3 - 1/G_1 G_2) - 1) - P_{sig,Rx,dBm} \right] / 20}. \end{aligned} \quad (17)$$

$$\begin{aligned} PWM_{Rx}(f) &= \sum_{n=-\infty}^{+\infty} A \frac{\sin \pi \frac{nW}{T}}{\pi \frac{n}{T}} \frac{\delta(f \pm \frac{n}{T})}{2} \Pi \left(\frac{f - f_{BW}}{2f_{BW}} \right) \\ &= \sum_{n=-M}^{+M} A \frac{\sin \pi n D}{\pi \frac{n}{T}} \frac{\delta(f \pm \frac{n}{T})}{2} \end{aligned} \quad (18)$$

where $f_{BW}T - 1 < M$ (integer) $f_{BW}T$, and A is the amplitude of the recovered pulses. To find the pulse width error due to the RBW limitation, we return (18) back to time domain using inverse Fourier transform

$$\text{PWM}_{R_x}(t) = \sum_{n=-M}^{+M} A \frac{\sin \pi n D}{\pi \frac{n}{T}} \cos \frac{2\pi n t}{T}. \quad (19)$$

By setting a threshold at $A/2$, and solving for $\text{PWM}_{R_x}(t) = A/2$ for t , we can find the recovered pulse width, $w_{R_x} = 2t$, and duty cycle $D_{R_x} = w_{R_x}/T$. It should be noted that if $M \rightarrow \infty$, i.e., unlimited RBW, then $2t = w$ and $D_{R_x} = D$. From (19) we have

$$\sum_{n=-M}^{+M} \frac{\sin \pi n D \cos \pi n D_{R_x}}{\pi \frac{n}{T}} = \frac{1}{2} = \sum_{n=-\infty}^{+\infty} \frac{\sin \pi n D \cos \pi n D}{\pi \frac{n}{T}}. \quad (20)$$

Subtracting $\sum_{n=-M}^{+M} \sin \pi n D \cos \pi n D / (\pi (n/T))$ from both sides yields

$$\sum_{n=-M}^{+M} \frac{\sin \pi n D (\cos \pi n D_{R_x} - \cos \pi n D)}{\pi \frac{n}{T}} = \sum_{|n| > M} \frac{\sin \pi n D \cos \pi n D}{\pi \frac{n}{T}}$$

which simplifies to

$$\sum_{n=-M}^{+M} \frac{\sin \pi n D \left(2 \sin \pi n \frac{D_{R_x} + D}{2} \cdot \sin \pi n \frac{D_{R_x} - D}{2} \right)}{\pi \frac{n}{T}} = \sum_{|n| > M} \frac{\sin 2\pi n D}{2\pi \frac{n}{T}}. \quad (21)$$

Defining $D_{R_x} - D = \Delta D$, and assuming $\pi n \Delta D \ll 1$, where $n \leq M$, then

$$\begin{aligned} \sum_{n=-M}^{+M} (\sin \pi n D) 2\Delta D &= \sum_{n=-M}^{+M} \frac{1 - \cos 2\pi n D}{2} \Delta D \\ &= \sum_{|n| > M} \frac{\sin 2\pi n D}{2\pi n}. \end{aligned} \quad (22)$$

Using (20)–(22), the RBW duty cycle error can be found from

$$\Delta D(M, D) = \frac{\sum_{|n| > M} \frac{\sin 2\pi n D}{2\pi n}}{\sum_{n=-M}^{+M} \frac{1 - \cos 2\pi n D}{2}}. \quad (23)$$

The numerator of (23) consists of the even terms of the PWM Fourier series in (5) that are left outside of the RBW and decreases with increasing the RBW (see Fig. 5). The denominator of (23) increases with M , which is proportional to RBW. Hence ΔD , decreases with increasing RBW. ΔD also depends on D and consequently on T as shown in Fig. 6. For a fixed sampling period, T , if D decreases (i.e., $W_2 < W_1$ in Fig. 6(a), the PWM spectrum spreads further out along with a reduction in the amplitude of its in-band components. Therefore, if the RBW is fixed, more power will be outside of the RBW and ΔD increases. A similar situation can occur if $D \rightarrow 1$. Because in that case, the denominator of (23) becomes very small, resulting in higher ΔD . This makes sense because when $D \rightarrow 1$ the

“low” pulses in PWM signal become quite narrow. Therefore, we need to limit $0 < D < 1$ from both ends.

Increasing the sampling rate without increasing the RBW can also have a detrimental effect on ΔD even if D is kept constant. This is demonstrated in Fig. 6(b), where sampling rate is doubled ($T_2 = T_1/2 \rightarrow M_2 = M_1/2$). This means that there will be more terms of the PWM spectrum outside RBW, which lead to a larger numerator in (23) and higher ΔD .

In Fig. 7, we have plotted ΔD versus D based on (23) for various RBWs from 5 to 75 MHz, and sampling rates of 320 and 640 kHz. In order to reduce ΔD , D has been limited to 10%–90%. Fig. 8 summarizes the theoretical worst case ΔD variations with RBW and D for 16-ch and 32-ch PWM-TDM systems, sampled at 20 kHz/ch. It can be seen that ΔD , which will eventually define the overall resolution of the WINeR system in wireless PWM technique can be adjusted based on the number of active channels and the desired sampling rate per channel. Therefore, unlike digital approaches, depending on the application, characteristics of the biological signal, and quality of the recording, the WINeR user can establish a tradeoff among the accuracy of the system, bandwidth per channel, and total number of active channels.

IV. Simulation and Measurement Results

In order to experimentally evaluate the performance of the wireless PWM technique on the WINeR system, we used the measurement setup, shown in Fig. 9. The 15-ch prototype WINeR ASIC, shown in Fig. 1(b), was fabricated in the AMI 0.5- μm std. CMOS process and wirebonded on a PGA 132-pin package. V_{high} , V_{low} , and I_C in Fig. 3 were adjusted for a total sampling rate of $f_{\text{PWM}} = 320$ kHz or 20 kHz/ch. The WINeR chip was battery powered at ± 1.5 V, consuming 4.5 mW, and the VCO was tuned to operate at $f_1/f_2 = 915/880$ MHz when receiving a rail to rail PWM signal. On the receiver side, we used a vector signal analyzer (Agilent 89600 VXI series) with tunable bandwidth from 5 to 36 MHz in addition to our custom designed wideband receiver with 75 MHz bandwidth [24]. The down converted IF-FSK-PWM signal was digitized at 4 GSample/s using an oscilloscope (Agilent MSO6104A) and the digitized data was further processed offline in a PC.

A. Matlab-Simulink Modeling

To further explore the effects of various system parameters on the wireless PWM performance, a realistic MATLAB-Simulink model was constructed including all the blocks shown in Fig. 1(a) and Fig. 2. Fig. 10 compares sample simulated versus measured waveforms when a ramp was applied to all 16 LNA inputs. In this experiment f_{LO} was tuned to 952 MHz, and after down-conversion, the IF-FSK-PWM signal was centered on 37 and 72 MHz [Fig. 5(a)]. The 37 MHz signal was located at the center of the receiver pass-band and receiver preserved most of its power, while the 72 MHz signal was located outside and was attenuated by the IF filters [Fig. 5(b)]. This signal can be easily envelope detected and sharpened by passing through a comparator to recover the PWM-TDM. Fig. 11(a) shows simulated samples of recovered PWM pulses when changing the RBW from 5 to 75 MHz.

In Fig. 11(b), we show the simulated results of duty cycle error from the 320 kHz MATLAB-Simulink model when RBW = 5, 10, 20, 36, and 75 MHz and $0.2 < D < 0.8$. According to these curves, which closely resemble theoretical outcomes shown in Fig. 7, with RBW = 75 MHz, the PWM technique can achieve $\Delta D < 10^{-4}$.

B. Measurements

Using the setup in Fig. 9, we experimentally evaluated the following wireless PWM performance measures.

1) Comparator Error—The PWM comparator error was evaluated by applying a dc voltage and a rail-to-rail sawtooth waveform from a precision function generator (Agilent 33250A) with 0.1% of peak linearity to the comparator inputs. The ideal pulse width when the sawtooth is greater than the dc voltage, and the actual output pulse width were compared for dc values swept from to -1.4 to 1.4 V in 0.1 V steps. The rms value for ΔD due to comparator error was found to be less than 7×10^{-4} .

2) TWG Error—According to (3), when V_{high} and V_{low} are constant, the TWG jitter is mainly due to the thermal noise of the CCSS and comparators. We sampled the TWG output at 2.5 GHz, subtracted it from a straight line with the same slope (ideal TWG waveform), and calculated the rms noise. The current noise is related to the voltage noise by a factor of C/T . Once the current noise is known, it can be used in (3) along with the comparator noise. Measurements showed that the rms ΔD from TWG was less than 10^{-3} .

3) VCO Error—Our measurements showed that the WINeR VCO ($f_{\text{osc}} \approx 900$ MHz) has a phase noise of S_{ϕ} (10 kHz) = -88.17 dBc/Hz. At sampling rate of $f_{\text{PWM}} = 320$ kHz and RBM = 36 MHz, $T_{\text{ffr}} = 74$ ns for the recovered PWM. Plugging these numbers in (11) results in $\Delta D \approx 10^{-5}$, which is far less than the other sources of error. Further, simulation and measurements showed that the VCO settling time is in the nanosecond range and it can be neglected with respect to T . Therefore, unlike high data rate digital wireless links, we can safely conclude that the VCO and LO on the transmitter and receiver units, respectively, do not have a dominant effect on the WINeR system accuracy.

4) Receiver Thermal Noise—The receiver noise figure is usually determined by the RF front-end LNA stages. Because they provide a high gain, which dominate the noise contribution by the following stages. Therefore, the entire receiver SNR is expected to be very close to the SNR at the RF LNA output. Our measurements at $d = 1$ m between the receiver and transmitter at the LNA output resulted in $P_{\text{sig}, Rx} = -15.7$ dBm and $P_{n, Rx} = -73$ dBm. Substituting these values in (17) yields $\Delta D = 1.4 \times 10^{-3}$. Even though this ΔD is not dominant in these measurement conditions, if the receiver SNR decreases as a result of increasing d or a strong interference, it has the potential to become the dominant source of noise and inaccuracy in the WINeR system. Proper matching between the receiver antenna and LNA input is also a crucial factor in improving the receiver sensitivity.

5) Receiver Bandwidth Limitation Error—For RBWs of 5 , 10 , 20 , 36 , and 75 MHz we directly compared the widths of the transmitted and received pulses, while changing the PWM duty cycle from 30% to 70% . Fig. 12 shows the measured ΔD versus D for $T = 3.125$ μs (320 kHz). It can be seen that RBW is one of the most effective parameters in defining the wireless PWM resolution, and it can dominate other sources of error if it is not wide enough. For the WINeR system operating with the custom designed receiver (75 MHz), the worst case ΔD was 2.4×10^{-3} at $D = 29\%$, which is roughly the equivalent of $-\ln_2(2.4 \times 10^{-3}) = 8.7$ bits of resolution. Considering the other sources of error discussed in Section III, we can conclude that the WINeR system can acquire 20 k-sample/s from 15 -ch plus the Mark signal with at least 8.4 bits of resolution. Table I summarizes the measured ΔD contribution from different sources of error across the entire WINeR system.

V. Discussion

In light of the introduction in Section I, and the fact that the majority of neuroscientists prefer to have access to the entire recorded neural signals across the wireless link as a seamless substitute to their hardwired data acquisition setups, the main alternative to our proposed TDM-PWM based architecture is the conventional digitization of the acquired

neural signals before wireless transmission [7]–[11]. In a $15 + 1$ channel system like the one shown in Fig. 1, the ADC should take 320 kS/s and produce 2.56 Mb/s of raw data, which has to be delivered across the wireless link. We have narrowed our comparison down to the following aspects of the system.

A. ADC Versus ATC

Both approaches need the multiplexer and S/H in Fig. 1(a). Hence the comparison narrows further down to the ADC architecture of choice versus Fig. 3, which is similar to the front-end of a single-slope ADC, in terms of speed, complexity, and power consumption [21]. To the best of our knowledge, there are few ADC architectures as simple as Fig. 3, which can offer the necessary performance. Moreover, the ATC method does not need any additional clock signal, which can be a potential source of substrate noise and interference with the on-chip LNAs [28].

1) Handling of the Information Before Transmission—As mentioned in Section II-A and shown in Fig. 1(a), the TDM-PWM signal directly drives the VCO. In an ADC-based architecture, on the other hand, digitized raw data has to be serialized, encoded, packetized, and combined with preamble and error detection bits. Even though these are routine tasks, the required digital circuitry can occupy a considerable chip area particularly in processes with large feature length, which are more suitable for low noise analog circuit blocks.

2) Fidelity of the Wireless Link—Low power wireless links such as Bluetooth 2.0 can offer data rates close to 2.56 Mb/s by relying on accurate time base generators that are crystal based. Crystals, however, cannot be integrated and occupy a large volume off chip. Transferring serial data at Mbps range with free-running VCOs does not seem to be feasible, and any effort in stabilizing the VCO frequency by using a phase-locked loop (PLL), for example, significantly increases the power consumption in the transmitter unit. On the other hand, the FSK-PWM-TDM signal has similar characteristics to digital FSK in being resistant to noise and interference. However, it does not need synchronization on a bit by bit basis. We showed that the VCO phase noise does not limit the WINeR system resolution, and its gradual drift can be compensated by tracking capability of the receiver.

Overall, we can conclude that unless there is a need for on chip DSP functions, as in the case of neuroprostheses, it is probably better not to digitize the neural signals on the transmitter unit. Our approach has been transferring the digital blocks and their associated area/power consumption outside of the body by dividing the ADC process into ATC on the transmitter unit and TDC on the receiver unit.

VI. Conclusion

Using a 15 channel wireless implantable neural recording (WINeR) system prototype, we have presented an effective architecture for simultaneously acquiring wideband neural signals from a large number of sites. WINeR operates based on pulse width modulation of time division multiplexed samples (PWM-TDM), which can potentially reduce the complexity, size, and power consumption of the implantable transmitter at the cost of adding to the complexity of the receiver without compromising the accuracy, robustness, or bandwidth of the entire system. It also provides the user with a high level of flexibility over the system resolution, sampling rate, and dynamic range. We have identified various sources of error in the proposed architecture as they affect the system resolution and analyzed their importance. We have supported our analysis with simulation and experimental measurement results. It turns out that the receiver bandwidth limitation is the dominant source of inaccuracy followed by SNR at the receiver RF front-end output. This information provides

useful guidelines in designing PWM-TDM based systems for a variety of wireless biomedical applications.

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Biographies

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Fig. 1.

(a) Block diagram of the transmitter unit in the wireless implantable neural recording (WIneR) system. (b) Die micrograph of the prototype WIneR ASIC with 15 channel neural signals and 1 channel for calibration, fabricated in the AMI 0.5- μm standard CMOS process [15].

Fig. 2.
Block diagram of the external receiver unit in the WINeR system.

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Fig. 4.
Transmitter FSK-PWM-TDM signal in time and frequency domains.

Fig. 5.
Receiver PWM-TDM signal in time and frequency domains.

Fig. 6.

Effects of RBW limitation on the spectrum of the recovered PWM signal. (a) Constant sampling frequency with reduced PWM duty cycle. (b) Constant PWM duty cycle with increased sampling rate.

Fig. 7.

Theoretical PWM duty cycle error due to receiver bandwidth limits at RBW = 5, 10, 20, 36, 75 MHz for (a) 320 kHz, (b) 640 kHz sampling rates.

Fig. 8.

Dependence of the PWM duty cycle error (ΔD) on sampling rate (f_{PWM}), PWM duty cycle (D), and RBW.

Fig. 9. Experimental setup for the wireless PWM technique using the WiNeR ASIC as the transmitter, and wideband receivers from 5 to 75 MHz.

Fig. 10.

(a) Simulated and (b) measured ramp input and PWM-TDM on the transmitter side, and IF-FSK-PWM signal on the receiver side.

Fig. 11.

(a) Simulated variations in the recovered PWM waveform due to RBW. (b) Simulated PWM duty cycle error versus D for different RBW.

Fig. 12. Measured wireless PWM duty cycle error versus D for different RBW at 320 kHz sampling rate.

TABLE I

Summary of Calculated Duty Cycle Error Contributions From Measured Parameters in the WINeR System

WINeR	Transmitter			Receiver	
Block	TWG	Comparator	VCO	RBW limitation	Receiver noise
ΔD	10^{-3}	7×10^{-4}	10^{-5}	2.4×10^{-3}	1.4×10^{-3}
Equivalent SNR	60 dB	63.1 dB	100 dB	52.4 dB	57.3 dB
Equivalent NOB	10.0	10.5	16.6	8.7	9.5
System Resolution	8.4 bits				

* $f_{PWM} = 320$ kHz, $f_1/f_2 = 915/880$ MHz, $0.3 < D < 0.7$, RBW = 75 MHz