A 0.5 cm³ Four-Channel 1.1 mW Wireless Biosignal Interface With 20 m Range

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Abstract—This paper presents a self-contained, single-chip biosignal monitoring system with wireless programmability and telemetry interface suitable for mainstream healthcare applications. The system consists of low-noise front end amplifiers, ADC, MICS/ISM transmitter and infrared programming capability to configure the state of the chip. An on-chip packetizer ensures easy pairing with standard off-the-shelf receivers. The chip is realized in the IBM 130 nm CMOS process with an area of $2 \times 2 \text{ mm}^2$. The entire system consumes 1.07 mW from a 1.2 V supply. It weighs 0.6 g including a zinc-air battery. The system has been extensively tested in in vivo biological experiments and requires minimal human interaction or calibration.

Index Terms—Brain computer interface, ECG/EKG, electrophysiology, EMG, low power, neuroscience, wireless.

I. INTRODUCTION

There is a tremendous emerging need for small, wireless, high-performance biosignal monitoring devices. There are existing devices that are currently deployed in relatively small numbers in animal research laboratories and serving as chronic implants. Researchers are also working on devices for clinical use and specialty athletic purposes. However, there are still barriers preventing widespread deployment of wireless devices in mainstream medical monitoring. Each application has specific challenges unique to its implementation, for example motion artifacts in body worn electrocardiograph (ECG) sensors and 60 Hz interference overriding low-level signal recordings. Each challenge can be addressed with a hardware approach on the device, but such a device becomes more application specific with these additions. The approach presented here takes a different view, preferring to create a device that is useful in a wide swath of applications, dealing with many of these application specific problems with either device programmability or digital signal conditioning off-chip. In this way we can develop a device that while it may not address each applications issues ideally, is useful for a wide range of diverse applications.

Our specification development was guided by the desire to have a device that worked over ECG, EMG, and spike-based neural recording paradigms. To define the needed technical requirements, two distinctly different applications were chosen that bookend the envisioned range of uses this device will be well suited for: human ECG and small animal (such as a mouse) extracellular single unit neuron recording. Human ECG exhibits fairly large voltage amplitudes (mVs) and low frequency content (150 Hz minimum). Single unit neuron recording voltage amplitudes can vary widely, from 10 s of $\mu$Vs to mVs depending on electrode placement and low frequency potentials (LFP). Frequency content here is much higher (over 1 kHz). While device weight is not a realistic limiting factor for a human, the device should be small enough and have a long enough wireless range to allow the subject freedom of movement. Conversely, weight and size are key constraints for applications on small animals such as a mouse. The device must be small enough for the animal to conformably carry without affecting its normal behavior. Both uses would benefit from a long lifetime running from a single battery; while in the ECG case its reasonable to extend the lifetime with a larger battery the small animal case requires a low-power design to maximize lifetime from a small battery due to the size constraint.

The main challenge lies in integrating the entire system in a small form factor of $< 1 \text{ cm}^3$, with a weight of $< 1 \text{ g}$ and low-power consumption ($< 10 \text{ mW}$) making them suitable for mass deployment. The signal chain in the system should have low input referred noise ($\text{few } \mu\text{Vrms}$) and sufficient dynamic range to accommodate the recording of different signal modalities (single unit recording, electrocardiography (ECG), electromyography (EMG)). Depending on the recording site location and condition the biosignal amplitude can vary widely, thus the gain and bandwidth of the amplifier front-end should be programmable on-the-fly to compensate. The monitoring device should also suppress motion artifacts to achieve long term portable biosignal monitoring. Lastly, the device needs to have a reliable, low-power communication interface to transmit the digitized biosignal to an external receiver. This necessitates the integration of a low-power transmitter that is capable of communicating reliably within an existing wireless infrastructure.

There has been a lot of recent effort to meet the above challenges. The configurability of the front-end depending on the recording mode and site requires closed-loop communication between the sensor node and the outside world. The wireless programming scheme should be simple enough to be used
by scientists and clinicians with little electronics background. Previous biosignal monitoring devices such as [1]–[5] fail to provide simple low power wireless configurability. Artifacts in recorded biosignals are common due to motion, bioelectrical activity, etc. If not mitigated these corrupting artifacts can seriously degrade the quality of recorded signal. Many techniques have been developed to combat these artifacts, for example the suppression of motion artifacts in the analog domain has been recently explored for ECG monitoring systems [6]. However, instead of implementing signal conditioning circuitry on-chip the system presented here post processes the digital signals from high dynamic range readout circuitry to tackle the problem of motion and other artifacts without biosignal specific circuitry, keeping the amplifier inputs more biosignal source agnostic.

The wireless telemetry of the digitized biosignals requires a low-power transmitter to establish a robust communication link with a basestation. The requirements of a wireless transmitter for a biosignal monitoring device include low power consumption, stable frequency generation, fast start-up time and compatibility with relevant frequency bands. The telemetry link could be established using an inductive link [7] or through RF communication. The range of inductive links is limited. Ultrawideband (UWB) is an attractive option, particularly for massively parallel multichannel systems where the data rate requirements are high [5]. However, this requires complex synchronization [8] and is susceptible to jamming due to the wide-open receiver RF front-end. Low power open-loop transmitters [4] experience drift as they are not locked to any reference signal using a crystal and hence their use in mainstream healthcare devices that require manufacturability and high yield/reliability is difficult. Therefore, a narrowband transmitter with a stable frequency generation scheme and low power consumption is a prudent choice for a wireless biosignal monitoring device.

Even though some of the challenges have been met individually, previous biosignal monitoring devices such as [1]–[4], [9] fail to achieve all of the stated objectives in a single system. This work represents the third generation of a single-chip system that allows robust wireless biosignal monitoring. Our goal is to describe decisions made in the design of the IC that allowed a low overall power dissipation and form-factor when deployed in actual monitoring environments. The paper is organized as follows. Section II outlines the system description with a focus on the Programming/Control Interface, the Analog Front End (AFE), the Analog-Transmitter Interface and the MICS compliant transmitter. Section III presents the measured results on the laboratory bench, in vivo experiments with mice and ECG monitoring of a human. Section IV concludes the paper.

II. SYSTEM DESCRIPTION

The overall architecture of the four channel biosignal monitor is shown in Fig. 1. The analog front-end comprises four low-noise, fully differential amplifiers designed to interface directly with electrodes connected to tissue, followed by a variable gain amplifier to accommodate different signal modalities. One amplified channel is selected and digitized using an on-chip 8-bit successive approximation register (SAR) ADC. The digitized data is buffered, appended with a programmable header, and tagged with cyclical redundancy check (CRC) information to detect communication errors. Finally, the digital bits are transmitted using a sub-mW frequency-multiplying transmitter. A programmable system clock is derived from the radio’s 48 MHz crystal reference. To enable on-the-fly programming of the chip, an infrared (IR) phototransistor interface is integrated into the system. The IR interface, tuned to 935 nm, allows wireless programming of the gain/bandwidth settings of each variable gain amplifier (VGA) independently and selects the active channel for transmission. All of these settings changes can be made on-the-fly using an infrared programmer to fine-tune the device for optimal recording during use.


**A. Programming/Control Interface**

A biosignal monitor often requires on-the-fly changes in gain and bandwidth to accommodate different recording site conditions and provide the ability to switch channels to target specific electrodes. A continuously running RF receiver would dissipate a significant amount of power, we instead propose a low power (<1 μW) wireless infrared interface using an off-chip IR phototransistor sensitive to 935 nm wavelength light. The incident IR signal is detected, amplified and limited rail-to-rail using an integrated comparator. A synthesized frequency detector block then measures the modulated frequency of the received IR signal. An IR remote control transmits the IR signal of a particular frequency for a fixed duration of 1 second. The digital frequency detector uses a divided version of the system clock and continuously counts the number of clock edges in a pre-programmed 333 ms time window to estimate the received signal frequency. If the determined frequencies of two successive time windows correspond to the same instruction then the signal is considered a valid instruction.

Once a valid instruction is detected the control block takes a corresponding action to change the state of the biosignal monitoring device as illustrated in Table I. A frequency-based encoding scheme was used to avoid the need for complex clock and data recovery to save power. Thus, this technique is limited to a relatively small number of commands; to mitigate this fact the number of programming commands was greatly reduced by implementing a control scheme where commands instruct the device to cycle between different settings as opposed to having many different commands that instruct the device to change to a particular setting. A frequency of 36 kHz increments the mux to select the next recording channel front end for digitization and transmission. Similarly, frequencies of 38 kHz and 40 kHz are used to cycle through the available gain and bandwidth settings. The sensor initially defaults into a reset state (with default gain and bandwidth settings hardwired through external switches) to suit the type of recording desired (neural, ECG, EMG, etc.). Current gain, filter and active channel settings are transmitted with each packet of sample data from the biosignal monitoring device so the user can confirm the device has been programmed as desired.

This implementation allows the device to be programmed from a few microwatts away while consuming negligible power in the IR receiver (a few microwatts). A tradeoff from using IR as opposed to various RF or near-field communications techniques is the line-of-sight requirement. IR light must have a fairly unobstructed path to the phototransistor to insure enough power arrives at the sensor for proper detection. For applications that cannot accommodate this, for example implantations, this receiving method is much less useful since the IR attenuation caused by tissue absorption would greatly reduce the programming distance, possibly enough to make programming impossible all together depending on implantation depth and tissue properties.

**B. Analog Front End**

The analog front end (Fig. 2) consists of four low-noise amplifiers whose inputs are AC coupled on-chip 20 pF capacitors. High-resistance (>100 GΩ) MOS-bipolar pseudo resistors enable a sub-Hz high pass cut-off frequency [10]. A complementary input stage in which both the n- and pFET’s of the input stage contribute to the effective transconductance [11] reduces the input-referred noise voltage by a factor of two for the same bias current [12]. A fully differential closed-loop architecture provides sufficient linearity and supply rejection. Thick oxide MOS transistors are used at the input to minimize the 1/f noise and reduce gate leakage. The VGA consists of a complementary rail-to-rail folded-cascode core with programmable capacitive feedback. Six-level variable gain is set by selecting the feedback capacitors while the seven variable high pass corners are set by programming the feedback transconductor bias current. Each of the four channels is individually configurable for different gain and bandwidth settings. This is useful in single channel deployments to select the best channel and then optimizing the AFE for the recording site characteristics. Table II illustrates the available gains and their high pass cut-off frequency range covered by the seven filter settings. Four complete analog front ends were included to allow fast switching between channels, not possible by muxing four inputs to a single AFE due to the IR front end design.
to the long settling time of the amplifiers compared to the sampling rate of the device.

C. ADC and Packetizer

An analog mux selects the amplified signal from one of the four input channels, which is digitized using a low-power 8-bit SAR ADC, designed to operate at sample rates from 10–100 kS/s. The SAR topology was chosen for the ADC to minimize power consumption [11]. Since the AFE settings are wirelessly programmable during operation the user can adjust the gain/bandwidth to provide a close to full swing signal to the ADC, allowing an 8-bit ADC to provide suitable dynamic range for various recording scenarios. The digital ADC output is read serially from the comparator output. A synchronization signal, which is used internally to purge the DAC capacitor array and SAR logic after each conversion, also synchronizes the serial output. A divider provides the baseband clock for the system from the 48 MHz clock generated in the local oscillator of the transmitter. The divide ratio is programmable to enable different sampling rates depending on the application.

A packet generator block (Fig. 3) samples the output of this ADC and prepares the data for transmission. The serial output data from the ADC is buffered in a first-in, first-out (FIFO) memory structure. Data is pulled out of the FIFO by the packetizer block, which inserts this data into a standard packet structure and delivers this packet to the transmitter. The packet header consists of 32 bits of alternating 1-0 s, then 32 bits of a constant sync code to identify the beginning of a packet, followed by 16 bits of system state data (current active channel number, gain and bandwidth setting), 256 bits of sample data and finally a 16 bit CRC. This packet structure was designed to be compatible with standard commercial ISM band receivers. The state data and sample data is run through a 16 bit CRC generator concurrently as the packetizer is outputting these bits. This synthesized digital logic block uses 2848 gates and takes approximately 0.018 mm² of area.

D. MICS/ISM-Compliant Transmitter

The RF transmitter is typically the most power hungry block in wireless bio-interface systems [13]. We address this by employing a low-power synthesizer architecture that operates entirely at the crystal reference of 48 MHz and drives a 9x frequency multiplying power amplifier (Fig. 4), eliminating the need for a PLL/DLL at the carrier frequency [11]. The baseband FSK data directly modulates the reference oscillator using capacitor pulling, allowing a 22 kHz frequency deviation. The reference clock drives a 9-stage DLL, thus the frequency deviation at the carrier frequency of 432 MHz is 198 kHz. This technique can be generalized to other multiplication factors, determined by the number of stages in the DLL and switching legs in the edge-combiner. Both DLL loops must demonstrate sufficient bandwidth to track the FSK baseband signal.

Matching between the delay stages is ensured by balancing the delay-stage loads, symmetric layout and dual-edge locking to minimize reference spurs. The edge-combiner behaves like a high-efficiency, non-linear power amplifier, and produces pulses of current based on overlap of separate delay cells in the DLL. This current is absorbed by a tapped-capacitor LC matching network, which transforms the TX source impedance to match a 50Ω antenna and attenuates out-of-band spurs. A return loss of less than −10 dB is achieved over the required bandwidth.

III. SYSTEM CHARACTERIZATION AND MEASUREMENT RESULTS

This IC was implemented in a 0.13 μm CMOS process, measuring (2 × 2) mm². Initial testing was performed on a large printed circuit board (PCB) to verify functionality. The die bonded to this test PCB can be seen in Fig. 5. The system is intended to accommodate two different battery chemistries: one with a nominal voltage around 1.2 V and another around 1.45 V. With this in mind, the chip was tested with both supplies. Running on a 1.2 V supply, 1.07 mW of power is dissipated by the chip, of which −15.1 dBm (30.9 µW) is transmitted as RF power. With a 1.45 V supply, 1.73 mW of
power is dissipated (RF output power of $-11$ dBm). Phase noise of the transmitter can be seen in Fig. 6.

### A. CMOS SoC

Rebooting into a relevant default state is key to this system’s ease of use. The IC restarts into a default running state; seven binary inputs are used to set the chip’s default AFE gain, bandwidth, and ADC sample rate. These settings are selected through a switch matrix of small SMT pull-up/down resistors on board allowing the same IC to default into appropriate AFE and sample rate states for different recording uses simply by changing these binary inputs. Of course, the chip parameters can be subsequently adjusted using the IR interface described in Section II-A. All settings function as expected.

### B. Deployable System

After benchtop verification, a small, lightweight PCB was designed to facilitate *in vivo* deployment of the system. The biosignal monitor was implemented on a four layer, $500 \mu m$ thick FR4 PCB measuring $(8.6 \times 9.7) \text{ mm}^2$, seen in Fig. 7. The top side of the PCB integrates the custom IC, seven 0201 pull-up/down resistor footprints to set different gain, bandwidth and clock speed options for the default state, a reset button, a $48 \text{ MHz}$ quartz crystal, an RF matching network, a $17.4 \text{ cm}$ long, No. 32 AWG insulated wire serving as a $1/4\lambda$ monopole antenna and an eight pin connector to mate external probes to the inputs of the device. Probes connect to this connector via a small flexible PCB that can be custom designed for the given experiment’s needs (i.e., single ended or fully differential, extra attenuation for large amplitude signals, integrated RC filter, etc.), allowing greater flexibility without requiring a re-spin of the biosignal monitor PCB. On the reverse side are either a 4.8 or $5.8 \text{ mm}$ diameter battery holder and an infrared phototransistor. The total mass of the device is dependent on which battery is used; loaded with the silver-oxide 337 battery the total system weighs $522 \text{ mg}$ including battery and antenna, with the zinc-air size 5 battery and a slightly larger battery holder the system weighs $612 \text{ mg}$.

The seven pull-up/down resistors could be removed to save area at the expense of fixing the default state of the device to only one option. Future versions could integrate non-volatile memory (NVM) to allow for different default states in the single design by programming the desired state into memory without requiring a reprogramming after every power-on. Likewise, future versions of this device could integrate a power-on reset block within the IC, eliminating the need for the external reset switch and reducing total deployable size and weight. The board area of the device is truly limited by the size of the battery and its holder; measuring $(6 \times 9) \text{ mm}^2$ for the zinc-air size 5 battery holder. Reducing the size any further would require the use of a different battery or an innovative mounting technique requiring a smaller footprint.
C. Receiver

A companion receiver for the sensor node was designed entirely of off-the-shelf components integrated on a (65×40) mm² PCB, as seen in Fig. 8. On board is a Texas Instruments CC430F5133 microcontroller with embedded wireless functionality, a USB interface translator to stream data to a PC, LEDs to indicate valid data reception, CRC failures and settings changes due to IR programming and a digital-to-analog converter (DAC) to view the reconstructed analog system in real time via an oscilloscope or audio output. Here we leverage the fact that the sensor node transmits a packet format compatible with this wireless chip, meaning all the build in data processing this chip is capable of can be used in our application, eliminating the need for custom components and greatly reducing the microcontroller code complexity. The receiver board is powered through a PC’s USB port so the current consumption of the receiver is not a critical constraint (around 30 mA in this case).

D. System Characterization

For system deployment, battery choice represents a critical tradeoff between size/weight and energy storage. The IC is capable of operating from a battery voltage anywhere between 0.9 V to 1.55 V. A variety of small coin cell type batteries are commercially available that could power this device, a selection of which can be seen in Table III. Of these, two batteries were chosen that best meet our needs of high energy storage density and a small form factor/low weight, the Zinc-Air Size 5 and the Silver-Oxide Size 337. In both cases the wireless sensor draws more current than these batteries were intended to supply, which limits the amount of energy the system can actually extract from either battery and slightly reduces the nominal voltages from their rated values. Fig. 9 shows the battery discharge curve for the Zinc-Air Size 5 battery as well as the received signal strength (RSSI) measured at our receiver board. This battery will power the device for over 22 hours with a fairly consistent voltage of approximately 1.2 V. Over the entire 22 hours of transmission there were 0 packet errors until the final few minutes of operation when the battery was almost completely exhausted.

The total power consumed by the sensor device is 1.07 mW using a 1.2 V supply (nominal voltage for Zinc-Air 5 battery) and 1.73 mW using a 1.45 V battery (nominal voltage for Silver-Oxide 337 battery). A key digital buffer on the IC was mistakenly powered from a diagnostic supply intended only for debug purposes. It was thus necessary to power this auxiliary supply continuously. This easily fixable design error contributes approximately 12% to the total power consumed by the device. Fig. 10 shows a breakdown of power consumption from the different blocks of the IC, including the auxiliary supply pin.

While it is possible to reduce the sampling rate of this device, the transmitter is in a fixed on-state, meaning a reduction in sampling rate will also reduce the baud rate of the transmitter to ensure a 100% duty cycle. Future versions could reduce the sampling rate while keeping the transmitter data rate at a maximum (200 kbps in this implementation), duty-cycling the transmitter as needed, which would greatly reduce power consumption for low sample rate applications. In this version, however, power consumption is fairly independent of sampling rate. Reducing
the sampling rate in the current device mainly serves to reduce the total amount of data the receiver/PC would need to process. Another important metric of any wireless system is the maximum acceptable range for data recovery. Commonly, this distance is defined as the point where the Bit Error Rate (BER, $P_b$) is $10^{-3}$. Packet Error Rate (PER, $P_p$) can be defined in terms of BER since the PER is simply the sum of probabilities of every possible packet error permutation for a given BER. This is shown in (1) & (2), where $n$ is the number of bits in a packet.

$$P_p(P_b) = \sum_{k=1}^{n} \left( \binom{n}{k} P_b^k (1-P_b)^{n-k} \right)$$  \hspace{1cm} (1)$$

$$P_p(P_b) = 1 - (1 - P_b)^n$$  \hspace{1cm} (2)

Our packets are 352 bits long; using (1) we see a PER of 29.7% represents a BER of $10^{-3}$. Fig. 11 shows the recorded PER of the system powered by the 337 Silver-Oxide battery at different distances from the receiver compared to the 29.7% PER point, giving a range of 24 m for this experiment. In a practical situation the acceptable amount of packet/data loss can be quite application dependent. Error detection is done solely by checking the CRC code so one cannot determine which bit(s) is/are in error in this system and thus must throw out the entire packet which makes the data loss in the received data “blocky.”

Fig. 11 can be used to decide the acceptable operating range of the system for a given application where practical range is decided by how many gaps in the recorded data is acceptable (likely over 20 m).

The total input referred noise of the chip was measured through the entire system shorting the analog inputs to ground and setting the AFE at the maximum gain (78 dB) and widest bandwidth (9.2 kHz) setting (Table II). The received data was then analyzed in Matlab to calculate input-referred noise. The wireless receiver recovered a signal with an RMS value of 1.73 $\mu$V and no packet errors. This RMS noise value was computed at the receiver and includes all noise sources, including amplifier noise, VGA noise, ADC quantization noise, aliasing, supply noise, substrate coupling, etc.

### IV. IN VIVO TESTING

#### A. Human ECG Recording

A key attribute of this system is its long lifetime (22 hours at 100% duty cycle) and ease of use. Operation requires only inserting a battery and pressing the reset button. This allows it to be easily deployed in a home setting that lacks any support equipment or technical staff commonly available in a laboratory setting that a less refined device might require.

To demonstrate this functionality, the sensor was attached to a freely moving human subject at home over the course of a day, recording the electrical activity of the heart (ECG). Two Ag/AgCl conductive adhesive electrodes were connected in a single-ended configuration to the wireless sensor and attached to the subject’s chest who was then left to go about his day while staying within range of the receiver. Fig. 12 presents a short length of this ECG data presented on a standard clinical ECG grid showing the expected healthy sinus rhythm. The signal was post-processed with a high order digital 1 Hz high pass filter to eliminate a wandering baseline, a changing dc level common in ECG monitoring likely due to patient movement.

An hour section of this recorded data was further processed, lining up each detected heartbeat with its central spike, known as the QRS complex. 4110 heartbeats were recorded in this hour, which can be seen overlaid in Fig. 13. The figure is in a heat-map format, allowing easy visualization of artifacts and heartbeat waveform morphology. The ECG data recorded from a freely moving subject is quite consistent over the length of the experiment, showing only modest changes in the t-wave, the later part of the ECG waveform, which is due to changing heart rhythm opposed to corrupted recordings.

Lastly, the recorded ECG data was used to determine the subject’s heart rate over two hours of normal activity, as seen in Fig. 14. There are minor variations in heart rate from beat to beat and over the two hour recording period. Here, 60 heartbeats were averaged together to calculate a mean heart rate for the subject and displayed in a more readable format.

#### B. Mouse Single Unit Recording

To demonstrate wireless neural recording from freely moving small animals, extracellular neural recordings from an awake, ambulatory mouse were performed. All surgical procedures were carried out in accordance with the Allen Institute for Brain Science’s Institutional Animal Care and Use Committee regulations using sterile techniques. Briefly, the cerebral cortex of anesthetized adult mice (C57/B16, 25 g) was implanted with...
a microwire electrode array, consisting of (4) 50 μm diameter polyimide-coated tungsten wires, [14], [15] coupled to the monitor via a custom flex board (Fig. 7). Electrical reference was attached using uninsulated tungsten wire connected to a bone screw in the left hemisphere. Electrodes were implanted into vibrissal somatosensory cortex, which can provide immediate spiking information to researchers via manual whisker deflection [16]. The mouse was allowed to recover from anesthetization before recording response from whisker stimulation, examples of which are shown in Fig. 15. Using the IR programmer the one electrode of four implanted closest to the cells of interest was exercised post-surgery by cycling through all four channels and recording from channels with active spiking. Likewise, the IR programmer was used to set the gain of this channel to utilize as much of the dynamic range of the ADC as possible without clipping the signal. This method of fine-tuning the recording session would be much more difficult with a wired or near-field communications system since the test subject is freely moving and is not easily held still to keep wires (or inductive coupler, etc.) attached.

With antenna and battery, the entire neural recording system weighed only 612 mg, and continuously streamed neuronal spiking data from freely moving mice for up to 22 hours. Such a system is capable of replacing large and complicated recording equipment, placing the technique of extracellular electrophysiological recordings into a wider variety of laboratories with limited space or resources. Data transmission was recorded at distances over 10 m with a USB-PC receiver interface.

Sensory-evoked neural spiking data recorded wirelessly were found to be qualitatively comparable to data recorded in a conventional wire-tethered recording system. At least two individual firing neurons could be identified from recordings in the somatosensory cortex of a freely moving, awake mouse, when this spike data was processed with sorting software [17]. These two spike waveforms, shown in Fig. 16, are displayed in a similar heat-map plotting technique to Fig. 13. Since these two neurons were firing at approximately equivalent intervals their waveforms interact and corrupt a single spike waveform. When we overlay all the spikes in this heat-map format however, the spike shape is clearly visible.

V. CONCLUSION

This paper demonstrates a low-power wireless biomedical sensor suitable for large-scale deployment in mainstream as well
as experimental medical applications. The system requires minimal human calibration, in the form of a reset switch to start the recording, and operates continuously for 22 hours from a single zinc-air battery. A receiver comprised of off-the-shelf components records the transmitted digitized data from distances of over 20 m. A simple IR wireless programmer allows amplifier settings to be changed during recording if needed to further fine-tune the setup. The entire biomedical wireless sensor consumes 1.07 mW from a 1.2 V supply.

The system’s ease of use allows it to be used in settings outside a laboratory and requires no technical knowledge to operate. To demonstrate this, the system was deployed in a home setting recording ECG and heart rate from a human over multiple hours without any need to tinker with the system. Highly configurable amplifiers make this system suitable for a wide range of recordings/experiments in addition to home monitoring. In addition to ECG recording, this system was used to record single neuron action potential firings within the brain of mice. The low weight and size of this device is beneficial for such small animal experiments, providing the experimenter with a way to record neural activity without weighing down the animal or confining it with a classical tethered system.

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REFERENCES


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