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A Wireless Magnetoresistive Sensing System for an Intraoral Tongue-Computer Interface

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Abstract

Tongue drive system (TDS) is a tongue-operated, minimally invasive, unobtrusive, and wireless assistive technology (AT) that infers users' intentions by detecting their voluntary tongue motion and translating them into user-defined commands. Here we present the new intraoral version of the TDS (iTDS), which has been implemented in the form of a dental retainer. The iTDS system-on-a-chip (SoC) features a configurable analog front-end (AFE) that reads the magnetic field variations inside the mouth from four 3-axial magnetoresistive sensors located at four corners of the iTDS printed circuit board (PCB). A dual-band transmitter (Tx) on the same chip operates at 27 and 432 MHz in the Industrial/Scientific/Medical (ISM) band to allow users to switch in the presence of

external interference. The Tx streams the digitized samples to a custom-designed TDS universal interface, built from commercial off-the-shelf (COTS) components, which delivers the iTDS data to other devices such as smartphones, personal computers (PC), and powered wheelchairs (PWC). Another key block on the iTDS SoC is the power management integrated circuit (PMIC), which provides individually regulated and duty-cycled 1.8 V supplies for sensors, AFE, Tx, and digital control blocks. The PMIC also charges a 50 mAh Li-ion battery with constant current up to 4.2 V, and recovers data and clock to update its configuration register through a 13.56 MHz inductive link. The iTDS SoC has been implemented in a 0.5- μ m standard CMOS process and consumes 3.7 mW on average.

Keywords

Assistive technologies; duty cycling; industrial-scientific-medical (ISM) band; intraoral Tongue Drive System (iTDS); magnetoresistive sensors; system-on-a-chip

I. Introduction

ASSISTIVE technologies (AT) are intended to enhance the independence of individuals with disabilities. Most ATs for people with physical disabilities try to focus on utilizing any remaining abilities of their users in the most efficient way to help them with activities of daily living (ADL) [1]. There has been considerable research toward the development of ATs by leveraging advancements in neurosciences, microelectronics, wireless communications, and computing [2]–[8]. Accordingly, there are many ATs available, which tend to enable individuals with severe disabilities to deliver their intentions to computers, as great equalizers that would provide them with the same benefits and opportunities that they offer to able-bodied individuals.

Even though a wide variety of ATs are available, many individuals with severe disability do not have a sufficient level of independence and still heavily depend on their family members or caregivers [9]. This is partly because the currently available ATs have limitations in one or more of the necessary aspects, which include intuitiveness, robustness, and compatibility [1], [10]. Clearly there is a need for ATs that are non- or minimally-invasive, easy to use, robust enough to operate well in various environments of daily life, and flexible and adaptive to offer broad coverage among potential end user populations as well as their temporary or permanent conditions, such as weakness, fatigue, and limited range of motion.

Tongue Drive System (TDS) is a minimally-invasive, wireless, and wearable tongue-computer interface (TCI) that offers multiple control functions over a variety of devices in the users' environments by taking advantage of their free tongue motion [11]. The tongue occupies a considerable area of motor cortex in humans, allowing it to be inherently capable of sophisticated motor control and manipulation tasks with many degrees of freedom [12], [13]. Individuals with the highest levels of spinal cord injury (SCI) can easily control their tongue with minimum thinking or concentration, and the tongue has a low rate of perceived exertion. Therefore, the tongue is a suitable choice as an intermediary connection to the brain to infer users' intentions, leading to development of several tongue-operated ATs [14]. A key advantage of the TDS is being wearable and always accessible to the users, allowing

them to switch between target devices without receiving help to switch from one dedicated AT to another.

TDS was originally developed in the form of a wireless headset, which we refer to as the external TDS (eTDS), using custom-designed circuits with commercial off-the-shelf (COTS) components, real-time magnetic sensor signal processing (SSP) algorithms, and a graphical user interface (GUI) [11], [15]. The eTDS prototypes were extensively evaluated on both able-bodied subjects and individuals with high-level SCI (C3-C6) at the Georgia Institute of Technology and Shepherd Center in Atlanta, GA, respectively, and the results were reported in [16]–[19]. These experiments proved the eTDS to be fully functional and quite effective in controlled environments, such as laboratory and hospital. However, the robustness of the eTDS in real-life indoor and outdoor conditions could not be guaranteed because of the poor mechanical stability of the headset. If there was a considerable shift in the sensor positions near the users' cheeks as a result of the eTDS headset displacement by an external force, *e.g.*, collision with an objects or driving in a bumpy terrain, the eTDS had to be recalibrated. Therefore, the eTDS had to be properly anchored to user's head using a headgear, which degraded its aesthetic aspects. These are problems that are shared with the most recent electro-encephalography (EEG) based brain-computer interfaces (BCI) that are implemented on headsets, mainly for entertainment applications [20]–[22].

To solve the mechanical stability issues and improve the TDS acceptability among its potential end users by making it an aesthetically transparent and highly reliable primary AT for their ADLs, we decided to anchor the magnetic sensors to the upper jaw, just below the palate, while hiding all other components inside the mouth after miniaturization. A similar approach had been previously adopted in a few tongue-operated ATs with various degrees of success [23]–[25]. In this paper we present the first intraoral TDS (iTDS) implementation. Most electronic components are embedded in a system-on-a-chip (SoC) that is the heart of a dental retainer, which tightly clasps onto the upper teeth to be well protected inside the oral cavity. Furthermore, unlike most other ATs, the iTDS will not be a sign of disability when it is completely hidden inside the mouth, offering its users a high degree of privacy [26].

Fig. 1 shows the overall block diagram of the new iTDS including its universal interface. The basic operating principle of the iTDS and its SSP algorithm are similar to the eTDS, which have been described in [15], [27], and will not be repeated here. The iTDS dental retainer wirelessly transmits (Tx) the digitized magnetic sensor raw data packets to a dual-band receiver (Rx) embedded in the TDS universal interface, which in turn delivers them to a smartphone, running the SSP algorithm, to be converted to user-defined tongue commands in real time. Unless the TDS commands are intended for the smartphone, they are sent back to the universal interface to be delivered to target devices, which can be a PC or powered wheelchair (PWC) in the current prototype [19].

The Section II explains the overall iTDS design considerations. Section III describes the architecture of the iTDS low-power mixed-signal application specific SoC, followed by the measurement results in Section IV. The overall system implementation is described in Section V, including the iTDS dental retainer and the TDS universal interface. Section VI

presents experimental results with a couple of able-bodied subjects, followed by the concluding remarks in Section VII.

II. iTDS Design Considerations

A. Design Choices

Using DC magnetic fields for wireless manipulation and displacement sensing is desired in medical devices because no power is needed for generating magnetic fields from permanent magnets. Yet another advantage of the DC and low frequency magnetic fields is that the human body is made of nonmagnetic material and completely transparent to such magnetic fields.

Magnetic field can be measured using a variety of methods from Hall-effect to flux-gate sensors, most of which have been implemented in small and low-power modules that include interfacing and compensation circuitry for positioning and mobile navigation applications [28]. Anisotropic magnetoresistance (AMR) sensors are arguably the top candidates for our TDS application because of their sensitivity range and sampling bandwidth, which are sufficient for tracking a small permanent magnetic tracer, attached to the tongue, within the 3D oral space.

For the iTDS, we adopted a 3-axial AMR-type magnetic sensor, HMC1043 (Honeywell, Morristown, NJ), because of its small footprint and low power consumption [29]. It contains three AMR Wheatstone bridges (one per orthogonal axis) for achieving high linearity, plus internal nulling mechanisms to cancel offset and drift. Four 3-axial HMC1043 sensors were placed on each corner of the printed circuit board (PCB) to provide sufficient coverage of the intraoral magnetic field variations. It can be shown that these 12 sensors can provide sufficient information for detecting the position and the orientation of a magnetic dipole tongue after cancelling the earth's magnetic field (EMF) interference [30].

Differential outputs of the HMC1043 bridges need to be amplified via precision instrumentation amplifiers to levels that would be sufficient for analog to digital conversion (ADC), which has a full scale input dynamic range of ground to $V_{DD} = 1.8$ V. Considering the HMC1043 sensitivity of 1 mV/V/Gauss and the strength of the field generated by the magnetic tracer at sensor locations (0.1–10 Gauss), four differential gain settings of 25, 50, 100, and 200 V/V were selected.

The iTDS sampling rate was determined considering the subject's reaction time, tongue dwelling time at each command position, and oversampling ratio of the magnetic sensor data. From our previous trials with the eTDS, the subjects' reaction time was 0.2–0.5 s, i.e., subjects maintained their tongues at command positions for 0.2–0.5 s before moving to another command position [13]. The magnetic sensor data need to be over-sampled at each command position to filter out the unintended tongue vibrations. We chose 50 Hz sampling rate for the eTDS to take at least 10 samples at each command position, which efficacy was empirically verified in human subject trials [18], [19]. In the case of iTDS, 64 Hz sampling rate was selected as the closest number to 50 Hz by dividing the 32.768 kHz iTDS system clock frequency by a factor of 512.

B. Design Challenges

To shrink the size of the TDS electronics to comfortably fit inside the mouth, the number of off-chip components and the size of battery should be minimized. An important design objective for the current prototype was to extend the continuous operating time of the iTDS beyond 24 hours, such that the users can effectively use it for two days (~ 12h/day) before a recharge would be necessary. Our solution was to integrate as many blocks as possible on a single chip (SoC) plus an adjustable power scheduling mechanism to duty-cycle the magnetic sensors, readout channels, and Tx to lower the average power consumption of the intraoral appliance [31].

Wireless data transmission from inside the mouth is another challenge because the RF carrier signal needs to pass through the oral and facial tissue before reaching the Rx. Human tissue is a strong attenuator of high frequency electromagnetic fields because of its water content (>90%). Table I shows the bulk characteristics and signal attenuation in the human muscle and fat tissue for the most popular industrial, scientific, and medical (ISM) bands [32], [33]. We chose 27 MHz as the default iTDS carrier frequency because it has the smallest attenuation. However, this band, which is also known as the citizen band radio, is also used for short-distance radio communications between individuals or remote controllers, and therefore, there is a relatively high probability of external interference.

To address this problem, the iTDS is equipped with a second low-power Tx that operates at 432 MHz. The user can switch between the two bands by altering the contents of the configuration register in the presence of a source of interference in the user's environment. The dual band transmitter improves the iTDS wireless link robustness by adding a certain degree of redundancy without consuming more power. Designing efficient miniature-sized antennas for each band is yet another design challenge, which has been discussed in Section IV-B.

The iTDS electronics should be hermetically sealed and carefully packaged in a dental retainer based on every user's unique oral anatomy to be safe and comfortable to wear over extended periods. Similar to implantable microelectronic devices, the hermetic sealing should prevent both leaching of moisture into the iTDS electronics and diffusion of harmful chemicals into the oral space. Proper shaping and fitting of the iTDS dental retainer can be accomplished using well developed orthodontic methods, starting from a mold that is prepared from the user's dental impression. However, since there is no way for making electric contacts to the hermetically sealed electronics, both recharging of the iTDS Li-Ion battery and programming of its configuration register should be accomplished wirelessly via an inductive link that operates at 13.56 MHz. A wire-wound coil, resonating at 13.56 MHz, has been embedded in the iTDS to receive AC power when it is placed inside the TDS universal interface. The same power carrier is amplitude shift keyed (ASK) for data transmission. The power management integrated circuit (PMIC) rectifies the received AC signal to charge the battery and amplitude demodulates it to recover the configuration data.

III. iTDS SoC Architecture

A. iTDS Building Blocks and Operation

Fig. 2 shows the block diagram of the iTDS dental retainer with more emphasis on the SoC architecture, which consists of these key building blocks: 1) an analog front-end (AFE) including three low-noise, high common mode rejection ratio (CMRR) readout channels operating in parallel with offset cancellation, 2) a low-power dual-band Tx operating at 27 and 432 MHz, 3) a PMIC, which includes wideband highly efficient rectifier operating at 13.56 MHz, a low-dropout voltage regulator, a battery charging circuit, and forward/backward data telemetry circuits, 4) multiple bias circuits and a reference generator, and 5) digital control circuitry for storing configuration settings and generating control signals for power scheduling various functional blocks for power saving.

To reduce the size and complexity of the digital control block, an ultra low power off-chip microcontroller (MSP430, TI, Dallas, TX) was utilized for data management. Since MSP430 already has an embedded analog to digital converter (ADC), it eliminates the need for such a block in the iTDS SoC. The magnetic sensors, AFE, Tx, bias/reference generator, and digital control blocks were supplied by 1.8 V regulated voltage, whereas the PMIC and analog switch control block were supplied at 2.8–4.2 V either directly from the battery, in normal operating mode, or from the on-chip rectified 13.56 MHz carrier when the battery was being inductively charged in the TDS universal interface.

Four 3-axial magnetic sensor modules are activated one at a time to measure the changes in the magnetic field at four corners of the palate. The three differential outputs of each sensor are time-division-multiplexed before connecting to the AFE inputs. Temperature data is sampled once every 64 samples and amplified through the same path as the magnetic samples. Three AFE outputs are multiplexed one more time towards the 10-bit ADC input of the MSP430, where digitized samples are also packetized. The dual-band Tx wirelessly transmits the digitized sensor packet to a COTS Rx inside the TDS universal interface, which delivers them to the smartphone for real time signal processing.

When the dental retainer is placed inside the charging cup of the TDS universal interface [see Fig. 13(a)], its embedded L_2C_2 tank in Fig. 2, tuned at 13.56 MHz, establishes an inductive link, which is driven by a radio frequency identification (RFID) reader, TRF7960 (TI, Dallas, TX). The RFID reader is capable of sending data to the iTDS by amplitude shift keying the carrier signal, which is both rectified and demodulated by the PMIC. The PMIC programs the iTDS configuration register and also echoes back the received data via back telemetry to validate the register contents. It also generates individually regulated 1.8 V supplies for various blocks and charges the 50 mAh Li-Ion battery up to 4.2 V. The configuration register consists of 6 bits which control the Tx carrier frequency and type of modulation, AFE gain, duty cycling ratio, and low pass filter cut-off frequency, as summarized in Table II.

The magnetoresistors on each leg of the HMC1043 sensing bridges change from 0.8 to 1.5 k Ω based on the magnetic field intensity and its orientation, leading to current drains of up to 2.25 mA per bridge at $V_{DD} = 1.8$ V. Obviously with such a high power consumption each

sensor module should be activated only as long as the AFE is sampling its output. Similarly, the Tx block has high power consumption and needs to be duty cycled. In our design, the duty cycling ratio of the magnetic sensors and AFE is selectable among 2.3%, 5%, 9%, and 18%, while the Tx is fixed at 25%. The duty cycling signal along with the sampling and time division multiplexing (TDM) control sequences are generated by the on-chip digital controller.

The ADC sampling times are always at the end of the sensors' on periods, as shown in Fig. 3, to minimize the AFE transient effects on the digitized samples. When selecting the aggressive 2.3% duty cycling ratio, the power scheduling mechanism turns one out of four magnetoresistive sensor modules and three AFE readout channels on for a total of 9.2% of a 15.62 ms sampling cycle, leading to 90.8% power saving, when added to 25% duty cycling of the Tx block.

B. Analog Front-End

Fig. 2 shows the AFE block diagram, including a 12×2 to 3×2 fully differential TDM multiplexer, three parallel readout channels (one for each sensor axis), a temperature-independent bias generator, and a single-ended sample-and-hold (S/H). In each AFE channel, the sensor bridge output is amplified by a fully differential current-feedback instrumentation amplifier (CFIA), providing an electronically selectable gain of 5, 10, 20 or 40 V/V, while achieving offset cancellation. Signals are then low-pass filtered by a linearized pseudo-RC filter, limiting broadband noise and aliasing, and then further amplified by a differential to single-ended amplifier presenting a fixed gain of 5 V/V, before being sampled at 64 Hz through a S/H amplifier featuring clock feed-through cancellation. Finally, the output samples from the three readout channels are multiplexed one more time and buffered before digitization in the MSP430. Details of the AFE design have been described in [34], and will be summarized in the following.

The first stage of the AFE, shown in Fig. 4, employs a CFIA with auto-zeroing to provide high input impedance, low noise, high linearity, high CMRR, low offset/drift, and precise and stable gain by means of isolation and balancing [35]. The circuit mainly consists of an input transconductance stage (G_{m1}) and a differential output transimpedance stage (R_m). A voltage follower (M_5, M_6) preceding G_{m1} accommodates with the common mode voltage at the magnetoresistive sensors' output (V_{ip}, V_{in}), which is $V_{DD}/2 = 900$ mV, while driving the gates of the $0.5\text{-}\mu\text{m}$ pMOS input devices (M_1, M_2) at a suitable DC level. The current through the input diff-pair (M_1, M_2) is held constant by a feedback loop consisting of G_{m2} and current sources $M_7 - M_{10}$. Feedback through G_{m2} balances G_{m1} and forces the differential input voltage to fall predominantly across R_1 , resulting in low-distortion. By mirroring the balancing currents onto the transimpedance stage, the circuit achieves high isolation, and the gain can be precisely set by the resistor ratio, R_2/R_1 , which is stable across process corners and temperatures. Moreover, opamp A_1 provides common mode feedback (CMFB) for the transimpedance stage.

Input devices have a large W/L ratio to achieve high transconductance with limited thermal noise, whereas $1/f$ noise has been reduced by employing devices with large $W \cdot L$ wherever possible. Offset cancellation was achieved through an auto-zeroing scheme, which consists

of sampling the output of the CFIA between each readout, after shorting the inputs by nMOS switches. Then, a correction current was applied in the transimpedance stage through feedback at the auxiliary port V_{os} , which drives the gate of M_{20} and maintain it until the next readout to balance this stage.

Aggressive duty cycling of the magnetoresistive sensors and the AFE results in considerable power saving. However, it also puts specific constraints on the AFE circuits. They must feature both fast settling and turn-on times, and also be robust against transients to prevent instability. Fast intermittent and yet accurate sampling requires the use of circuits with larger bandwidth, which let in more noise. These trade-offs are made in selecting circuit topologies for the iTDS SoC along with programmable duty cycling and tunable filtering, as shown in Fig. 5. The cutoff frequency of the tunable low-pass filter was selected with respect to the duty-cycling employed in the system, according to Table II, to yield a suitable transient response while filtering broadband noise. Such an arrangement provides different levels of accuracy and battery lifetime depending on the task at hand and the iTDS user preference.

Programmable low-pass filters (LPF) have been designed in the past by changing the gate voltage of pseudo-resistors [36]. These methods are susceptible to process variations and suffer from distortion when the input signals are large. In Fig. 5, we have utilized linearized pseudo-resistors with V_i -independent V_{GS} and a group of small capacitors that were electronically selected from a bank to achieve stable and high-linearity cut-off frequencies for different duty cycles, as shown in Table II, which are also tunable against process variations by changing V_{ctrl} . In this design, non-inverting amplifiers along with resistor voltage dividers ($R_1 = 500 \text{ k}\Omega$) reflect any variations in the large input and common mode voltages onto the gates of the pseudo-resistors, thus preventing nonlinearity, while providing an adequate DC level to set the desired cut-off frequency.

Simulations showed that in order to achieve a total harmonic distortion (THD) of $<0.1\%$, the LPF bandwidth should be higher than 2 kHz at 2.3% duty cycle. A 0.5 kHz error margin was considered in the design of the LPF at this duty cycle, which corresponds to $V_{ctrl} = 17 \text{ mV}$ when taking into account the $0.5 \text{ M}\Omega/\text{mV}$ sensitivity of the $30 \text{ M}\Omega$ pseudo-resistors to V_{ctrl} . Therefore, $V_{ctrl} = 900 \text{ mV}$ was generated from V_{DD} with 2% accuracy ($\sim 17 \text{ mV}/900 \text{ mV}$) using an external resistive voltage divider with 1% accurate surface mount resistors.

Differential output of the LPF is fed into a differential amplifier with a fixed gain of 5 ($R_3 = 5R_2$) while being converted to a single-ended signal before sampling. Four poly-resistors employed in this circuit are carefully matched for achieving precise gain and high CMRR. It is because a slight mismatch of the differential gain may cause unnecessary common mode output which distorts the differential amplifier output. The S/H circuit employs a non-inverting configuration with clock feed-through cancellation that prompts for very-low THD. When in track mode, the nMOS switches, S_1 and S_2 , are closed and the high gain and low output impedance opamp is connected in unity gain configuration. When going into hold mode, the nMOS switches open, and the input voltage is sampled. Note that all analog switches employed in the AFE block are implemented using minimum feature size nMOS devices, with gates controlled by a high voltage level (2.8–4.2 V) taken from the battery, through the PMIC block, allowing high linearity and low charge injection.

C. Dual-Band Wireless Transmitter

Fig. 6 shows a simplified schematic diagram of the dual-band low power Tx. Two crystal oscillators, operating at 27 and 48 MHz, with identical circuits have been utilized to generate accurate timebase, using off-chip crystals. The 432 MHz carrier is generated by multiplying the 48 MHz clock by 9, utilizing a dual-loop delay locked loop (DLL) and an edge combiner [37]. The 27 MHz Tx includes a class-C PA preceded by a 27 MHz crystal oscillator, as shown in Fig. 6.

The DLL-based topology for the 432 MHz Tx is attractive because of its lower power consumption, better stability, less jitter, and faster locking time compared to phase locked loops (PLL), in which a power hungry voltage-controlled oscillator (VCO) is needed to create the high frequency carrier. The edge combiner behaves like a nonlinear power amplifier and injects sufficient current to the off-chip LC-tank, which matches the 50 Ω antenna to the Tx output. The dual-loop DLL includes 9 voltage-controlled delay cells, which have been designed with current-starved inverters, phase-frequency detectors (PFDs), and charge pumps. Two identical feedback loops delay the rising- and falling-edges of the 48 MHz clock in each cell to generate nine 50% duty cycle, equally spaced clocks.

Schematic of the dynamic PFD, which is used to control the charge pump, is shown in Fig. 7 [38]. The PFD can be divided into two halves that are identical except for input signals that are switched. Each block consists of two cascaded stages with a pre-charge pMOS in each stage. Dynamic PFD eliminates flip-flops in conventional PFDs, which dissipate high power at higher clock frequencies, while maintaining a simple structure and fast transition times. The specific charge pump topology that is shown in Fig. 8 was chosen to reduce charge injection, clock feed-through, and charge sharing in conventional charge pumps. The source and sink currents are carefully matched by increasing the length of current mirror transistors, M_1 and M_2 , and adding a replica branch (T_3 and T_4 switches) of the charge/discharge branch (T_1 and T_2 switches) to avoid switching of the current source transistors.

In the FSK mode, serial data directly modulates the reference frequencies using crystal pulling, as shown in Fig. 6. The OOK mode has been implemented by turning on/off the buffers before the class-C PA or the edge combiner in the 27 and 432 MHz Tx blocks, respectively.

D. PMIC With Bidirectional Data Telemetry

During normal iTDS operation, the PMIC is only in charge of power scheduling and bias generation, thus most of its sub-blocks are off. However, when the iTDS dental retainer is placed inside the charging cup of the TDS universal interface [see Fig. 13(a)], the 13.56 MHz power carrier couples onto the L_2C_2 tank, generating an AC signal across the full-wave active rectifier inputs, which supplies the rest of the PMIC and charges the iTDS embedded 50 mAh Li-Ion battery, as shown in Fig. 9. The design and operation of the active full-wave rectifier, which offers high AC-to-DC power conversion efficiency (PCE) in the order of 80% at 13.56 MHz thanks to its offset-controlled high speed comparators and optimally sized switches, can be found in [39].

The PMIC has bidirectional data telemetry capability with the RFID reader in the TDS universal interface that drives the inductive link. Fig. 10 shows the schematic diagrams of the clock and data recovery circuits for the forward data telemetry. Clock recovery circuit in Fig. 10(a) generates the clock signal by comparing the 13.56 MHz sinusoidal signal across the L_2C_2 tank. The recovered clock is then buffered and divided by 256 to provide a 53 kHz master clock signal for the rest of the system. For data recovery, variations on the V_{REC} due to ASK of the power carrier by the RFID reader are fed into the data recovery circuit in Fig. 10(b). This simple circuit detects V_{REC} amplitude variations using two paths with different time constants, $R_1C_1 < R_2C_2$, which are connected to a hysteresis comparator. The difference between input node voltages following V_{REC} amplitude transitions results in the recovered forward data bit stream at the output of the comparator, which are sampled and delivered to the configuration register by the back telemetry controller, as shown in Fig. 9. This circuit also generates a short pulse for every detected bit “1” and applies it to the load-shift-keying (LSK) mechanism of the active rectifier [39]. Shorting the rectifier input results in a sudden drop in V_{IN} and increased current in L_1 . The current and voltage variations in L_1 are detected by the RFID reader and used to recover the LSK back telemetry data.

A low dropout regulator (LDO) follows the rectifier, which provides the battery charger block with a constant 4.4 V supply. Battery charger provides a constant charging current of 6.8 mA to the battery as long as $V_{BAT} < 4.2$ V. When V_{BAT} is charged near 4.2 V, the charger switches from constant current to constant voltage mode and keeps V_{BAT} at 4.2 V to continue charging the battery without damaging it. During constant voltage mode, the charging current gradually decreases, until the charger stops charging the battery when the current goes below 5% of its nominal value.

Once V_{BAT} reaches its maximum charging voltage of 4.2 V or the inductive link powering is removed, the battery monitoring circuit disables the battery charger operation as well as connects the battery to the system supply for starting the normal iTDS operation.

The PMIC has been equipped with a detuning-based over-voltage protection circuit, which compares $V_{REC}/4$ with V_{REF} , as shown in Fig. 9, and closes a switch that detunes the L_2C_2 tank when V_{REC} is too high. Detuning is a safety measure that results in a considerable drop in V_{IN} , which prevents possible damage to the active rectifier and other circuits when the rectifier output voltage has grown too large as a result of the coils being too close or the load current being too small.

IV. System Implementation

A. iTDS Dental Retainer

Fig. 11 shows the first implementation of the iTDS in the form of a prototype dental retainer. Electronics are supported on a trapezoidal 4-layer FR4 PCB that easily fits inside the mouth of an average adult. Nonetheless the size of the current iTDS prototype is bulkier than a regular dental retainer. As such, it can limit the range of tongue movements and may not be comfortable for long term use. In the future we plan to shrink the size of the iTDS dental retainer to comfortably fit inside the mouth by further reducing its power consumption, which can lead to shrinking the size of the battery, eliminating the test circuits

on the PCB, more integration, and using a much thinner Polyimide-based flexboard instead of the rigid FR4 PCB.

Once the board passes functionality tests, it is hermetically sealed with a thin layer of Parylene-C or medical-grade epoxy. The sealed board is then placed inside a mold that is built based on the user's dental impression before being potted with a self-curing acrylic resin, which is commonly used in fabrication of orthodontic appliances. Additional metallic arches and ball clasps were added to the dental retainer to help fix it onto the upper teeth, as shown in Fig. 1.

B. Dual-Band Antennas

The iTDS needs to operate inside the mouth and aside from the insulating materials. Its antenna is surrounded by a complex time-variable oral environment that includes air, saliva, teeth, gum, bone, facial and tongue muscles, and skin tissue, which considerably affects the RF signal. Moreover, the available space is far below the optimal dimensions at the designated iTDS frequencies. Design of the antenna and matching circuit for the iTDS is particularly challenging because they should handle not only the attenuation caused by the electromagnetic energy absorption in the surrounding nonhomogeneous tissue, but also the variable impedance resulted from tongue movements and jaw positions [40]. What is presented here is an early attempt in addressing the above issues. As such there is room for considerable improvements in this aspect.

To save space, the stainless-steel orthodontic arch wire was used as the 27 MHz antenna with one end connected to the feed point of the 27 MHz Tx output and the other end open. The antenna was matched with a wideband L-matching network because the antenna wire will be in direct contact with the gum. The capacitive loading improves the antenna's response because of its electrically small length. This also makes the antenna more robust against the complex and variable oral environment. For the 432 MHz band, a monopole spiral antenna was designed with numerical optimization with a tapped-capacitor LC matching network to the 432 MHz Tx circuit. Fig. 12 shows the 3D simulation model and actual prototypes of both antennas, which are explained in Section V-D.

C. TDS Universal Interface

The custom-designed TDS universal interface serves several important purposes in the overall functionality and usage of the iTDS dental retainer on a regular basis: 1) Dual-band receiver (Rx) at 27 and 432 MHz for the raw magnetic sensor data from the iTDS; 2) Inductive charger of the iTDS at 13.56 MHz when the dental retainer is placed in the special charging cup. The same inductive link can update the iTDS configuration register via ASK and validate it via LSK; 3) Holding the smartphone within the field of view of the iTDS end-user either mounted on a PWC or simply on a flat surface; 4) Deliver the received raw data from iTDS to the smartphone for applying the SSP algorithm and detecting user commands; 5) Establish a reliable hardwired interface between the smartphone and PWC to convert and deliver the TDS navigation commands in the form of DC-level adjusted analog signals [19].

TDS universal interface was encased in a custom enclosure, shown in Fig. 13(a), which was designed in SolidWorks (Waltham, MA) and fabricated using 3D rapid prototyping

technology. Its internal COTS circuitry, the block diagram is shown in Fig. 1, has been implemented in 3 floors and a vertical coil, as shown in Fig. 13(b). The heart of the interface is an MSP430 SoC microcontroller that manages other blocks and communicates with the smartphone (iPhone), computer, and PWC via 30-pin, USB, and D-type 9-pin (DB9) connectors, respectively. A power management (PM) block on the 1st floor utilizes three input power sources: 12 V from PWC, 5 V for USB, and 3.7 V from a built-in 2000 mAh Li-Ion rechargeable battery located on the 3rd floor, with priority given to the highest available voltage source. Regardless of the input source, the PM block generates two main supplies, 5 V and 3 V, using buck-boost DC-DC converters and voltage regulators. The PM block also charges the interface battery from the PWC or USB input power.

The 2nd floor houses the RF circuits including the dual-band COTS Rx and its 27 MHz whip antenna connector and 432 MHz chip antenna and their $50\ \Omega$ matching circuits. We used a CC1110 (TI, Dallas, TX) transceiver for 432 MHz but the 27 MHz Rx was custom-designed using discrete components. The 13.56 MHz inductive link for charging and programming was driven by a TRF7960 RFID reader, as mentioned in Section I, which primary L_1C_1 tank PCB was vertically positioned behind the iTDS charging cup, as shown in Fig. 13.

V. Measurement Results

The iTDS SoC was fabricated in the ON-Semi $0.5\text{-}\mu\text{m}$ 3M2P standard CMOS process, resulting in a $3.8 \times 3.7\text{ mm}^2$ chip, which micrograph and floor-plan are shown in Fig. 14. Fig. 15(a) presents the measured sensor supply waveforms with and without duty cycling and Fig. 15(b) represents the power consumption break out in each case. It can be seen that the total power consumption at $V_{DD} = 1.8\text{ V}$ drops by 71.8% from 13.2 mW without duty cycling when each sensor is turned on for a quarter of the sampling period to only 3.7 mW with aggressive duty cycling when each sensor is turned on for 2.3% of the sampling period. In addition, the AFE and Tx blocks are also turned off for 90.2% and 75% of the period, respectively.

A. Analog Front-End

Table III summarizes the measurement results of the AFE. The AFE consumes $473\ \mu\text{A}$ current under $1.8\text{ V } V_{DD}$ without duty cycling. It achieves THD of 0.06% at 100 Hz with a gain of 100 V/V. The CMRR and the PSRR were measured as 84 dB and 56 dB, respectively. Fig. 16 presents the multiplexed output waveform of the AFE when the gain is set to 100 V/V. The AFE amplifies 12 differential analog inputs from four 3-axial magnetic sensors and multiplexes them at 1 kHz sampling clock. The AFE output varies from 0.3 to 1.5 V according to the magnetic field strength. Each output is sampled at the falling edge of the ADC sampling clock by the external MSP430 microprocessor.

B. Dual-Band Wireless Transmitter

Fig. 17 shows the measurement results for the dual-band Tx when operating in the OOK mode at 64 kbps data rate. Fig. 17(a) shows the serial data bit stream applied to the 432 MHz Tx and the resulting output carrier signal across a $50\ \Omega$ terminated load. Fig. 17(b)

shows the frequency spectrum of the 432 MHz OOK carrier, which achieves a maximum output power of -15 dBm while consuming 1.13 mA from a $V_{DD} = 1.8$ V. The measured BER in air was $<10^{-6}$ when using a 5 cm monopole Tx antenna, 4 m away from the Rx. In the FSK mode, the crystal oscillator's frequency varied by 13 kHz, leading to 117 kHz carrier frequency variation around 432 MHz.

The 27 MHz Tx output power was set at -6 , considering that more external interference was anticipated in the 27 MHz band. Fig. 17(c) waveforms from top show the serial data bit stream applied to the Tx input, the resulting OOK carrier signal, the power detector output on the Rx sides when the iTDS was placed in the mouth and held ~ 30 away from the Rx whip antenna, shown in Fig. 13(a). The power detector output was high pass filtered and compared with a reference voltage to detect the serial data bit stream, shown in the bottom waveform of Fig. 17(c). Fig. 17(d) shows the frequency spectrum of the transmitted 27 MHz OOK signal. The 27 MHz Tx consumes 1.35 mA from $V_{DD} = 1.8$ V. In the FSK mode, the measured crystal oscillators' frequency deviation was ± 13 kHz.

C. PMIC With Bidirectional Data Telemetry

Fig. 18 shows the measured PMIC results. In Fig. 18(a), the active rectifier receives 13.56 MHz sinusoidal waveform at $V_{IN1} - V_{IN2} = 5.9$ V_{peak} from L_2C_2 tank and converts it to 4.6 V DC output. The LDO generates a 4.4 V regulated output that supplies the rest of the PMIC. The measured PCE of the active rectifier was $\sim 75\%$ at 6.9 mA load current, 6.8 mA of which was dedicated to the battery charger. Fig. 18(b) shows the battery voltage and the charging current profile. For $V_{BAT} < 4.2$ V, the battery is charged up at 6.8 mA. When V_{BAT} nears 4.2 V, the battery charger provides a constant voltage of 4.2 V, while the charging current gradually drops. The 50 mAh Li-Ion battery takes ~ 8 hours to be fully charged through the inductive link.

In Fig. 18(c), the clock recovery circuit converts the 13.56 MHz inductive carrier to a 13.56 MHz clock signal, which is divided by 256 to generate the 53 kHz clock signal that is used by the rest of the iTDS SoC. Fig. 18(d) shows the ASK demodulator waveforms. From top, the 24 V_{p-p} sinusoid across the primary L_1C_1 tank is ASK-modulated by the RFID reader with a modulation index of 33% , which appears across the secondary L_2C_2 tank and V_{REC} on the 2nd and 3rd traces, respectively. Finally, the comparator in Fig. 10(b) recovers the serial data bit stream at 1 kbps on the bottom waveform.

D. Antenna Radiation Patterns

There have been prior attempts in characterizing intraoral antennas [40], [41]. However, they were either focused only on 2.4 GHz or the antenna position and type were different from the ones used in the iTDS. There are various sources of signal loss in the wireless channel between the intraoral Tx and the external Rx, as shown in Fig. 19, which are quite difficult to predict accurately in this sub- λ distance that is considered near-field [42]. To simplify the analysis and validate the antenna design, we characterized the 27 and 432 MHz wireless links, including the overall link losses and antenna radiation patterns, in both horizontal and vertical planes.

In simulation, the output power of the iTDS was applied to the Tx antenna and received power was calculated at the Rx antenna using the CST Microwave Studio (Framingham, MA) and Zubal Phantom human body model, shown in Fig. 12(a) [43], [44]. Moreover, the extracted S_{11} data from the iTDS board was applied to the simulations in order to consider the imperfect matching loss (L_M) at the Tx antenna. Thus, simulation results take into account the human body attenuation (L_H) and matching conditions. Tx antenna gain (G_{Tx}) and air attenuation (L_{AIR}) are also considered by the simulator based on the antenna geometry and Tx-Rx distance, which was set to 30", (~ 75 cm) as the nominal distance between the iTDS and TDS universal interface. The Rx antenna gain (G_{Rx}) was considered ideal, meaning that all the power arrived at the Rx antenna was delivered to the Rx input.

All measurements were performed in an open space, shown in Fig. 20(a) and (b), to avoid multipath from reflections. All measurements were performed with the iTDS dental retainer in the subject's mouth, as shown in Fig. 1, with the mouth closed. Standard high gain commercial antennas were used to measure the antenna patterns, with ideal matching to the Rx. Fig. 20(c) and (d) show the simulated and measured radiation patterns at 27 MHz in the horizontal and vertical planes, respectively. Fig. 20(e) and (f) show similar results for the 432 MHz link. These measurement results match simulations in terms of both shapes of radiation pattern and amount of attenuation. The minimum received power levels were -68 and -77 at 27 and 432 MHz, respectively, with the input power levels of -11.8 and -21.5 dBm. Considering that the sensitivity of the COTS receivers at 27 and 432 MHz is in the order of -90 to -100 dBm, the FSK/OOK-modulated data from the iTDS should be successfully recoverable with this range of received signal power, unless there is a strong in-band interference [45], [46].

VI. System Performance Evaluation

To operate the iTDS, users need to attach a small magnet on the tongue using tissue adhesives (temporarily) or piercing (semi-permanently) [19]. After attaching the magnet and wearing the iTDS dental retainer, the system needs to be calibrated and trained [11]. Calibration is used to cancel out the effects of EMF [15]. The iTDS is then trained with 7 different tongue positions corresponding to 6 TDS commands plus neutral, which is the tongue resting position. Now, when users position their tongues in those pre-defined positions, the system detects and executes the issued command in real time.

A. Experimental Setup for Computer Access

Two healthy subjects (one male and one female) aged 32 and 28 years old, who were among the TDS development team, participated in this experiment, which included a computer access task using the setup shown in Fig. 21. Both subjects were familiar with the eTDS and had participated in our prior human subject trials, but they were new to the iTDS [17]–[19]. Two iTDS dental retainers were built by a professional dental technician based on the subjects' dental impressions. Subjects wore the iTDS retainers, attached magnetic tracers to their tongues, sat ~ 1 m away from a 22" LCD monitor, and then calibrated and trained the iTDS. A TDS universal interface received the sensor data from the iTDS and delivered it to a desktop PC via USB, as shown in Fig. 21. Fig. 22(a) shows the training results for one of

the subjects, indicating that the command clusters were separated enough to be easily distinguished from each other by the SSP algorithm.

Maze navigation task was selected to quantitatively evaluate the iTDS performance in navigating the mouse cursor in an emulated environment. Subjects were instructed to move the cursor from the “START” to the “END” circle by following the blue track in Fig. 21 as quickly and accurately as possible, using their tongue commands. Both subjects repeated the maze navigation task for 4 rounds (one practice and 3 testing rounds), and successfully completed them.

Two performance metrics were used: deviation from the path and completion time, which are indicators of the accuracy and speed of the subjects and the computer access device [18]. Deviation measure was calculated by integrating all the outside area between the designated path and actual traversed path of the mouse cursor, as shown in Fig. 22(b). The average values of the deviation and completion time using the iTDS were $4.8 \pm 2.0 \text{ pixel}^2/1000$ and $12.1 \pm 2.7 \text{ s}$, respectively.

B. Comparison Between iTDS and eTDS

To compare the performance of the iTDS with the eTDS, which has already been evaluated not only by able-bodied subjects but also in clinical trials [16], the same subjects were asked to perform the aforementioned task using the eTDS headset, as shown in Fig. 21. The average deviation from path and completion time with the eTDS were $4.27 \pm 0.84 \text{ pixel}^2/1000$ and $11.5 \pm 1.05 \text{ s}$, respectively. These results, which are depicted in Fig. 23, are slightly better than the iTDS both in terms of absolute value and less variability.

The reasons could be: 1) iTDS occupies some of the intraoral space and limits the tongue movements. This issue was more pronounced in the current iTDS prototype, which had been implemented on regular 1.6 mm thick rigid FR4 PCB with several additional off-chip components for characterization, as opposed to thin and flexible polyimide; 2) the position of the current iTDS on the roof of the mouth blocks tactile feedback that the users would otherwise receive from their palates, which can result in better positioning of the tongue tip; 3) both subjects had prior experience with the eTDS but were new to the iTDS; 4) subjects were stationary and did not experience any inertial or mechanical interference.

On the other hand, the authors expect that in a more extensive trial involving PWC navigation over a longer period of time, the iTDS can produce better result than the eTDS because: 1) The iTDS dental retainer is customized to the users' oral anatomy and it firmly clasps onto their teeth. Therefore, sensor displacement, which can occur to the eTDS when the headset shifts as a result of strong inertial or mechanical input, is absent. Better stability for the sensor positions with respect to the magnetic tracer over longer periods results in more stable operation and less demand for command re-training. 2) Signal-to-noise ratio (SNR) is considerably higher in the iTDS because of the proximity of the sensors to the magnetic tracer. This will provide the iTDS with more robustness against electromagnetic interference and particularly the EMF. 3) Future revisions of the iTDS dental retainer will be smaller, lighter, and can be shaped like a horseshoe to place the sensors and electronics over the outer rim of the upper or lower teeth in order not to block the intraoral tactile feedback.

C. Benchmarking

Table IV compares the iTDS specifications with other tongue operated ATs from literature. Both the ITCS and Optical Tongue Gesture device are intraoral [24], [25]. However, unlike iTDS, their 1D sensors cannot detect the tongue position within the 3D oral space. As a result, their tongue commands are neither adjustable nor adaptive, and bound to be located at specific positions inside the user's mouth.

VII. Conclusion

A new intraoral TDS (iTDS) has been presented as a complete system, including an ultra low power SoC and a COTS TDS universal interface. The high precision AFE collects analog samples from 12 magnetoresistive sensors and delivers them to an off-chip microcontroller for digitization and formation of uniform data packets, which are transmitted by an on-chip dual-band Tx, operating in the ISM-band (27 and 432 MHz). The PMIC block reduces the power consumption to 3.7 mW with aggressive duty cycling and wirelessly charges an embedded Li-Ion battery for continuous 24 h operation. The universal interface wirelessly receives the raw data packets and delivers them to a smartphone for translation to tongue commands, which are then delivered to PC or PWC. The first iTDS prototypes were implemented in the form of dental retainers and their performance was verified via a computer access test, and also compared with the eTDS. Key advantages of the iTDS are considered to be its mechanical stability, robustness against noise and EMF interference due to its high SNR, and being completely hidden inside the mouth. There is room for considerable improvements in the iTDS wireless connectivity using body area network (BAN) and intra-body communication techniques. We are also working on real time adaptive matching circuits and improved antenna geometries to deal with the dynamics of the oral space.

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Biography

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Fig. 1. Overall system block diagram of the intraoral Tongue Drive System (iTDS) including a magnetic tracer, a dental retainer, and a universal interface.

Fig. 2. Block diagram of the iTDS SoC application specific integrated circuit (ASIC), which is the heart of the iTDS dental retainer.

Fig. 3.
Timing diagram of the AFE and ADC with adjustable duty cycling.

Fig. 4. Schematic diagram of the current feedback instrumentation amplifier (CFIA) with offset cancellation, used at the input of the iTDS AFE.

Fig. 5. Schematic diagram of the pseudo LPF, differential amplifier, and sample-and-hold (S/H) circuits.

Fig. 6.

Schematic diagram of the dual-band Tx that can operate at 27 MHz or 432 MHz and transmit data by applying either OOK or FSK.

Fig. 7.

Schematic diagram of the dynamic phase-frequency detector (PFD) used in the 432 MHz Tx to reduce power consumption at 48 MHz.

Fig. 8.

Schematic diagram of the charge pump used in the 432 MHz Tx to reduce the charge injection, charge sharing, and clock feed-through.

Fig. 9.

Schematic diagram of the power management IC, including rectifier, regulator, battery charger, and bidirectional data telemetry.

Fig. 10. Schematic diagram of (a) clock recovery and (b) ASK data recovery circuits for forward data telemetry.

Fig. 11.
Actual implementation of the iTDS in a form of a dental retainer.

Fig. 12.

(a) Simulation model. (b) Actual implementation of the 27 MHz and the 432 MHz iTDS Tx antennas.

Fig. 13.

TDS universal interface implementation with 3D rapid prototyping. (a) Showing the covered iTDS charging cup, 27 MHz Rx antenna, smartphone holding functions, and hardwired connection to a PWC. (b) Internal view of the interface boards including an a 13.56 MHz charging coil, PMIC board, RF board and a 2000 mAh rechargeable battery.

Fig. 14.
Chip micrograph and floor-plan of the iTDS SoC in ON-Semi 0.5- μm .

Fig. 15. Measurement result of (a) the sensor supply voltages and (b) the resulting power consumption with and without duty cycling.

Fig. 16.

Measured waveform of the multiplexed AFE block output. The ADC in MSP430 samples this signal at the falling edges of the sampling clock.

Fig. 17.

Dual-band Tx measurement results in the OOK mode with 64 kbps serial data. (a) Serial data bit stream and 432 MHz carrier at Tx output across a $50\ \Omega$ load. (b) Frequency spectrum of the signal in (a). (c) Serial data bit stream and 27 MHz carrier at Tx output across a 50 load, Rx power detector output and recovered data bit stream. (d) Frequency spectrum of the signal in (c).

Fig. 18.

(a) Measured waveforms of the active rectifier and LDO inputs and outputs. (b) Li-Ion battery inductive charging profile, showing its switching from constant current to constant voltage mode at ~4.2 V. Measured waveforms of the (c) clock and (d) the data recovery circuits within the PMIC block.

Fig. 19.
RF link losses of the path from the iTDS Tx to the COTS Rx.

Fig. 20.

The measurement setup for the antenna radiation pattern and channel loss in (Φ) horizontal rotation, and (b) vertical (θ) planes. Measured (solid line) and simulated (dashed line) radiation patterns of (c) 27 MHz antenna in horizontal plane, (d) 27 MHz antenna in vertical plane, (e) 432 MHz antenna in horizontal plane, and (f) 432 MHz antenna in vertical plane.

Fig. 21.

Test setup for the maze navigation task with both iTDS dental retainer and eTDS headset.
 Inset: 6 recommended command positions plus neutral.

Fig. 22.

(a) Training result with the iTDS mapped onto the 3D PCA space. (b) The actual cursor movement trajectory for one of the subjects, used to measure the deviation from the designated path in the maze navigation task.

Fig. 23.

Performance comparison between the eTDS and the iTDS. (a) Sum of deviation from the path. (b) Completion time for the maze navigation task.

TABLE I

Frequency-Dependent Attenuation Characteristics of Tissue [33]

Carrier Frequency (MHz)	Conductivity (σ , S/m)	Relative permittivity (ϵ_r)	Attenuation coefficient (α , Np/m)	Attenuation per cm (dB)
27.12	0.70 */0.03 **	120/10	7.60/1.07	0.33/0.06
433.92	1.00/0.05	60/5	23.1/3.73	1.00/0.16
915	1.40/0.06	55/4.5	34.6/4.85	1.50/0.21
2450	2.50/0.09	50/4.2	65.5/8.07	2.85/0.35

* Muscle/

** Fat

TABLE II

Configuration Register Functions

Bits	Description
1	Carrier frequency selection: (0, 1 = 27 MHz, 432 MHz)
2	Modulation scheme: (0, 1 = OOK, FSK)
3, 4	AFE gain selection: (00, 01, 10, 11 = 25, 50, 100, 200)
5, 6	Duty cycling selection: (00, 01, 10, 11 = 2.3, 5, 9, 18 %) Low-pass filter cutoff: (00, 01, 10, 11 = 2500, 1000, 500, 250 Hz)

TABLE III

Performance Summary of the Analog Front-End

Performance parameter	Value
Supply voltage (V)	1.8
Current consumption without duty cycling (μA)	473
THD% @ 100 Hz input with 100 V/V gain	0.06
Common mode rejection ratio (CMRR) up to 3 kHz (dB)	84
Power supply rejection ratio (PSRR) up to 3 kHz (dB)	56

TABLE IV

Summary of the iTDS Specifications and Benchmarking

Specifications		iTDS	eTDS [11]	ITCS [24]	Optical Tongue Gesture [25]
Process		0.5- μ m std. CMOS	COTS	COTS	COTS
Die area		$3.8 \times 3.7 \text{ mm}^2$	-	-	-
VDD		1.8 (V)	3.0 (V)	-	-
Sensor	Type	Magnetoresistive	Magnetoresistive	Inductive-coupling	Infrared-proximity
	Channel	12	12	18	4
	Sensitivity	1.8 (mV/V/gauss)	3 (mV/V/gauss)	-	0.2-0.5 (V/mm)
	Sampling	64 (Hz)	50 (Hz)	30 (Hz)	90 (Hz)
RF	Frequency	27/432 (MHz)	2.4 (GHz)	2.4 (GHz)	Hardwired
	Data rate	64 (kbps)	500 (kbps)	-	
Power	Consumption	3.7 (mW)	15.6 (mW)	-	Hardwired
	Battery	Li-ion 50 mAh	Ni-Mh 800 mAh	-	
	Lifetime	24.3 (hour)	123 (hour)	15 (hour)	
Prototype	Dimensions	$49 \times 42 \times 15 \text{ (mm}^3\text{)}$	$23 \times 20 \times 19 \text{ (cm}^3\text{)}$	$35 \times 25 \times 15 \text{ (mm}^3\text{)}$	$40 \times 40 \times 15 \text{ (mm}^3\text{)}^*$
	Volume	20 (ml)	69 (ml)	-	-
	Weight	75 (g)	170 (g)	-	-
Number of commands		6	6	6	4
Applications		Computer access / PWC navigation	Computer access / PWC navigation	Computer access / PWC navigation	Computer access

* Estimated from publication.