Energy Efficient Capacitive Body Channel Access Schemes
for Internet of Bodies

Thesis by
Abeer AlAmoudi

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The thesis of Abeer AlAmoudi is approved by the examination committee

Committee Chairperson: Ahmed Eltawil
Committee Members: Khaled Salama, Basem Shihada
ABSTRACT

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Abeer AlAmoudi

The Internet of Bodies (IoB) is a wireless network of on-body or in-body communication formed by wearable, ingestible, injectable, and implantable smart devices. The vast majority of on-body communications, is typically required to be within <5 cm vicinity of the human body. The radiative nature of currently used RF devices leads to wasted energy that is radiated in unneeded off-body directions. Consequently, it degrades energy efficiency, introduces co-existence and interference problems, and imposes security threats on sensitive data. As an alternative, the capacitive body channel communication (BCC) couples the signal (between 10 kHZ-100 MHZ) to the human body, which is more conductive than air. Hence, it provides lower loss, better privacy and confidentiality, and nJ/bit to pJ/bit energy efficiency. Accordingly, our work investigates orthogonal and non-orthogonal capacitive body channel access schemes for ultralow-power IoB networks with or without cooperation. We derive the closed-form optimal power allocation for uplink and downlink transmissions and the maximum number of IoB nodes satisfying a reliable and feasible network for non-cooperative schemes. The cooperative schemes necessitate joint optimization of both power and phase time allocations. We achieve this by using the Golden-Section search algorithm to minimize the power consumption in both phases.
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Chapter 1

Introduction

1.1 Motivation and Background

In today’s climate, following the COVID-19 pandemic, the necessity of movement towards a scalable, cost-effective, and smart health care system has become more apparent than ever. Hospitals and healthcare centers across the globe, with their lack of readiness in managing an unpredictable large-scale scenario, experienced a surge in COVID-19 critical cases. Thus, causing a dramatic depletion in resources and situating many on the brink of collapse. As of June 1, 2021, there have been 170 million confirmed COVID-19 infections and around 3.5 million deaths [2]. These numbers are expected to go higher with the ongoing COVID-19 spread. Therefore, transforming the infrastructure of the medical sector is of the utmost importance to manage the current situation and better prepared for future crises. In this work, we delve into the Internet of Bodies (IoB) and its diverse applications towards this goal.

The Internet of Things (IoT) is a novel paradigm revolutionizing multiple sectors and industries through the integration of physical and digital realities [3]. Hence, it is owed to advancements in wireless communication and Micro Electro Mechanical Systems (MEMS). The integration of sensing, communication, and processing capabilities in devices leverages the connectivity and interaction between multiple entities [4]. An imminent derivation of IoT is IoB aimed at human body digitization. In a nutshell, a wireless network of on-body or in-body communication is formed by wearable, ingestible, injectable, and implantable smart devices [1].
The IoB, which was initially addressed under wireless body area networks (WBANs) [5], captures the essence of the widespread and interconnectivity of IoT. Thus it has a broad range of potential applications. In this context, patient monitoring, preventive healthcare, wellness, fitness, and cybersecurity are few of the IoB based applications. We refer interested readers to Fig.1 in [1] for a holistic view of the different applications. Further, the importance of IoB in healthcare-based applications is amplified with the current world health issue COVID-19. Hence, the constant aggregation and analysis of physiological data and vital signs have positive outcomes in detecting new cases, providing real-time monitoring to self-quarantined patients, and preventing infections [1]. However, exploiting the full potential of IoB is contingent on addressing the interrelationship between its requirements and design challenges and associated risks.

1.2 IoB Requirements and Design Considerations

While most of the IoB experimental setups are readily implemented with the aid of existing technologies, however, the aim for robustness, security, and scalability calls for further research. Given the broad range of applications, the IoB system design is subject to trade-offs in “QoS, safety, power efficiency, form-factor, and privacy “requirements. In particular, the quality-of-service design metrics (QoS), which include throughput, reliability, latency, bit error rate (BER), and energy efficiency, are determined by the underlying application. For instance, applications demanding high QoS urge for sophistication in hardware design and processing capabilities. This infers an increase in the power demand, hence, opposing the objective of ultralow-power IoB node and leading to a shortened battery life. Moreover, on account of commercializing IoB, security and confidentiality concerns arise due to the collection/transmission of sensitive data over the air, which necessitates the use of different security measures to prevent malicious attacks and the disclosure of user privacy, such as key generation
and agreement protocols. However, the implementation of such protocols is limited by the inherent power, storage, computational, and QoS constraints [6]. Lastly, since the IoB includes localization and communication of devices in-on-and-around the human body, the foremost priority is to ensure that IoB design complies with health and safety guidelines. The primary health concerns include heating and radiation effects on the body. Therefore, entities like the International Commission on Non-Ionizing Radiation Protection (ICNIRP) propose limits on the exposure to electromagnetic (EM) radiations. This is measured by specific absorption rate (SAR) in [W/Kg], and in turn, SAR defines the maximum transmission power to avoid tissue heating, and damages [7].

1.3 Communication Standards

For the successful deployment and generalization of IoB, communication and networking standards are required to address its various requirements. In this regard, multiple WBAN standards, including IEEE 802.15.4, IEEE 802.15.6, and Bluetooth LE, are referenced to define the bases for future IoB standards. Both IEEE 802.15.4 and Bluetooth LE operating on industrial-scientific-and-medical (ISM) bands suffer from coexistence and interference due to the ever-growing number of IoT devices operating on these bands. Further, both standards failed to satisfy high data rates, and low power consumptions [7]. Later the IEEE 802.15.6 standard was established to resolve limitations of IEEE 802.15.4 and Bluetooth LE and meet the QoS metrics [8].

The IEEE 802.15.6 standard provides three physical (PHY) options narrowband (NB) and ultra-wideband (UWB) radio frequency (RF) communications and body channel communications (BCC). Referring to [1], it is evident that from channel modeling of NB and UWB, the signal attenuation in, on, and around the human body varies significantly. Which is attributed to the heterogeneous frequency-dependent dielectric nature of the human body. Although RF-based systems are widely employed
because of their level of maturity, yet in the context of IoB, they are inadequate for on-body communications. Since on-body communication typically is required to be within <5 cm vicinity of the body. However, the radiative nature of currently used RF devices leads to wasted energy radiated in unneeded off-body directions. Consequently, it degrades energy efficiency, introduces co-existence and interference problems, and imposes security threats on sensitive data. Moreover, the power-hungry radio-front ends of RF-based devices calls for larger battery capacity and continuous battery charging, hence, reduces the battery lifetime and causes inconvenience for the user, and impedes commercialization.

Alternatively, BCC, designated to operate between 100 kHz and 100 MHz, couples the signal to the human body having higher conductivity than air. Thereby, BCC is subject to a lower path-loss, translating into less transmission power requirements \[9\] \[10\]. Another key attribute for the endowment of IoB in BCC is its operational frequency for two main reasons. Firstly, frequencies below 100 kHz are susceptible to (EM) interferences, and secondly, at frequencies beyond 100 MHz, the body act as an antenna \[11\]. Moreover, operating at below 100 MHz enables carrier less communication. Thus, the transceiver size can be decoupled from the carrier wavelength, eliminating the need for radio front-ends. Lastly, this operational frequency reduces signal leakage \[12\], enhances energy efficiency and physical layer security in turn. Overall, IoB can leverage from the promising technology BCC is set to deliver, hence, ultralow-power, miniature, and low-cost devices with better physical layer security attributes.

1.4 Thesis Contributions and Objective

To date, most of the research efforts are directed towards channel modeling \[1\] and references therein] optimizing transceiver designs \[13\] \[14\] \[15\]. However, integrating these valuable contributions with networking and communication aspects is yet to
be studied. Accordingly, in this thesis, we take a step forward towards enabling IoB by proposing an energy-efficient regular and cooperative orthogonal (OMA) and non-orthogonal (NOMA) capacitive body channel access schemes for ultralow power IoB network. To this aim, the main contributions of this thesis can be summarized in the following points:

• Present closed-form optimal power allocation for uplink (UL) and downlink (DL) transmissions. In comparison with numerical solvers, closed-form solution illustrated 32% better fairness for a cancellation error ($\epsilon=1^{-9}$) and Channel length ration 3, which enhances the IoB network lifetime substantially.

• For a given bandwidth, we obtain the maximum number of IoB nodes satisfying a reliable and feasible network for non-cooperative schemes for two scenarios:

1. Subject to different QoS requirements: The obtained results for both UL and DL displayed the inverse relation between number of users and higher QoS requirements. For UL traffic, NOMA outperforms OMA by 72% for QoS 0.5 Mbps if node 1 is 20 cm away from the hub. Further, if node 1 is at 100 cm distance from the hub, then NOMA can accommodate 22% more users than OMA for QoS 0.5 Mbps. For DL traffic it is shown that OMA accommodates 60% and 55% of what NOMA is capable of for node 1 at 20 cm and 100 cm respectively.

2. Subject to varying the transmission power requirements: We also found a positive relation between transmission power and number of nodes for UL and DL. In UL traffic for $P = -40$ dBm, the maximum number of nodes in OMA is 72% and 57% of NOMA’s maximum capacity for node 1 at 20 cm and 100 cm respectively. In DL for $P= -40$ dBm, OMA can realize 78% and 60% of what NOMA hosts for node 1 at 20 cm and 100 cm respectively.

• Present the network lifetime as a function of energy departure rate and QoS
constraints, and emphasizing the effect of transceiver energy efficiency.

- Present cooperative and cooperative schemes and model their performance:

1. Investigate the impact of the relay node’s channel gain, hence, its location with respect to the hub node on the energy consumption.

2. Effect of the source nodes’ channel gains, hence, their location with respect to the relay node on the effective transmission energy.

3. Address the impact of the network’s size on both orthogonal and non-orthogonal cooperative schemes.
Chapter 2

Literature Review

2.1 Coupling Methods

The development of BCC is owed to the extensive research efforts dedicated to understanding the behavior of human tissue in the presence of electromagnetic (EM) fields. Thus, providing fundamentals of signal propagation mechanisms and channel characteristics. In BCC, the transmitter electrodes confine the communication signals to the body by means of electrostatic or magneto-static fields, which are captured at the receiver end. Hence, there are three main coupling mechanisms to implement BCC: capacitive coupling, galvanic coupling, and magnetic coupling [1]. In what follows, the details will be presented.

2.1.1 Capacitive coupling

Capacitive coupling (CC), also referred to as near-field coupling and electrostatic coupling, was first introduced by Zimmerman in [16] under personal area network (PAN). Initially, in the electric field sensing experiment, it was observed that electric signals distort when a hand is placed inside the field. This suggests that the human body can be utilized to convey digital information using capacitive coupling. The setup of CC is presented in Fig. 2.1, includes signal electrodes of both transmitter and receiver, which are in contact with the skin, leaving the ground electrodes to float in the air. A stimulus electric field is then applied to the signal electrode at the transmitter end. As a result of the body’s conductivity, the electric signal is captured
in the body. Hence, it creates a forward path while the environment and earth ground serve as a return path [17]. Consequently, the conveyed signal in BCC is subject to external effects such as interference from nearby fields. Nevertheless, CC-BCC is a promising mechanism set to deliver high QoS demands for low transmission power levels.

Figure 2.1: Capacitive Coupling based BCC [1]

2.1.2 Galvanic coupling

Galvanic coupling was proposed by Handa et al. in [18] for a wireless ECG monitoring system. The ECG signals gathered from the detector mounted on the subject’s chest were transmitted to a wrist-worn electrode by passing $\mu$A AC current through the body. Not limited to ECG monitoring, the GC-BCC can be deployed for other applications as long as both signal and ground plates of both transmitter and receiver are touching the skin as illustrated in Fig. 2.2. Hence, the transmitted signals are confined within the body, granting GC-BCC immunity to environmental effects. However, since the operational frequency of GC-BCC is $< 1$ MHz, it is incapable of supporting high-throughput and long-range communications [17].
2.1.3 Magnetic coupling

In both CC-BCC and GC-BCC, the electro quasi-static fields are employed to facilitate the communication between the transmitter and receiver. In contrast, the magnetic coupling MC-BCC, suggested in [19], utilizes magneto quasi-static field to reach longer distances while mitigating environmental effects. As shown in Fig. 2.3 a magnetic field (in red dashed lines) is generated by the current (in yellow dashed lines) carrying coil. Moreover, an effective loop is invoked when both foot and hand, on the coil side, simultaneously touch the ground and signal electrodes, respectively. Hence, the channel is disconnected by detaching the hand, foot, or both from the electrodes. Further, the robustness of MC-BCC was validated in [19] where no significant attenuation to the signal was observed in the presence of metal objects.
Given the abovementioned, it is in our best interest to establish a network based on a coupling mechanism that would provide high gains and throughputs at high frequencies while requiring low power control levels. Hence, we will focus our discussion in the subsequent sections and chapters on CC-BCC, which is supported by IEEE 802.15.6 standard.

2.2 Physical Layer

2.2.1 Channel Modelling

A thorough understanding of the channel characteristics serves as the backbone for the development of effective BCC systems. Further, accurate channel modeling is vital to device efficient BCC transceivers [20] which are the main components in the communication system. However, this task is obscured by the uncertainties induced by various factors, including environmental effects, electrode placement on the body, varying human tissue structure and impedances, channel length, and varying electrode specifications [19] [21] [17]. In this context, this section will present the main channel modeling techniques reported in the literature.
Analytical Models

The analytical model is developed to cultivate a basic understanding of the EM propagation within and around the body. The EM characteristics for simplified geometries, which emulate body parts, are derived by solving Maxwell’s equations for specific boundary conditions. In [22] the authors proposed a model for CC-BCC consisting of quasi-static near-field, inductive field, and the surface far-field. This model was intended to study the interrelation between path loss, frequency, and channel length. The model proved to be valid for frequencies 100 kHz-100 MHz and up to 1.3 m distance. However, considering simple geometries in analytical models does not avail the designing process of a practical BCC system. Because it ignores the various tissue compositions leading to a discrepancy in theoretical and experimental results [21].

Numerical Models

The intricate anatomy of the human tissues necessitates the use of numerical computations to provide accurate channel modeling. In the literature, finite element method (FEM) and finite difference time domain (FDTD) are the most adopted numerical methods to investigate the dielectric properties of the tissue and the optimal electrode configuration that would strengthen the signal [21].

The finite element method (FEM) is a rigorous tool that analyzes electromagnetic structures by dividing the spatial domain into smaller elements, usually triangular or tetrahedral-shaped. The solution is then approximated from these discretized elements by solving a set of linear equations. Hence, the complexity of the problem subject to the number of unknowns there are. Which is controlled by the size of the discrete grid [23]. In [24] FEM is used for CC-BCC, the body was modeled by a $\pi$-shaped circuit to look into the body influence on the return path. The results obtained demonstrated that below 4 MHz, the channel gain is separate from the forward path.
and is controlled by the capacitive return path. However, beyond 4 MHz, impedances of the forward path and leakage from the body to the ground distort the signal.

The finite-difference time-domain (FDTD) is a powerful numerical tool intended for computational electromagnetic modeling problems. The method proposed by K. Yee in 1966 [25] solves Maxwell’s differential equations directly in the time domain. As a result, a single simulation will unleash the system’s behavior for a wide range of frequencies. Hence, its prominence among other numerical methods is on account of its reduced computational complexity and its ability to analyze complex structures [23]. One of the most significant applications of FDTD is determining the human exposure to electromagnetic radiations, i.e., SAR [26]. In [27] and [28] the FDTD was employed for BCC to obtain the distribution of electric field around the body. The authors concluded that the electric field is mostly centralized on the tip and surface of the arm, i.e., the field is confined in the body.

There are many efforts for applying FEM and FDTD in BCC modeling, we refer interested readers to [1, and references therein]. Although numerical methods achieve higher accuracy than analytical methods, they operate at the expense of time and computational cost. Thus, implying the need for moderate models to deliver precision and adequate running time.

**Circuit Models**

Circuit models provide simplified yet accurate solutions with better processing capabilities for broad frequency ranges [1]. These simplified models account for the electrical attributes of the different body parts, such as their resistive and capacitive nature and frequency-dependent dielectric properties [20] [21]. Hence, providing insights on the signal attenuation and dispersion [21]. Circuit elements of bio-impedances represent the body. Then mathematically interpreted as transfer functions describing the propagation of the signal from the transmitter to the receiver through the body.
Mainly, BCC circuit models are classified under lumped circuit parameters because, below 100 MHz, the physical size of the body parts is small when compared to the signal’s wavelength [1].

Accordingly, in [20] they represent a five-layer tissue lumped circuit model for capacitive coupling. This model enhances precision because it accounts for the surface physiological properties of the body, and it analyses the five layers with lumped elements. Thus, it can adapt to diverse body compositions in different subjects. Moreover, this model is compatible with IEEE 802.15.6 standard having a central frequency of 21 MHz, wherein the suggested model and measurements are set for a frequency range of 1 MHz-40 MHz. Nonetheless, there are other approaches where distributed circuit parameters are utilized to mitigate signal loss experienced at the receiver due to high transmission lengths and high-frequency requirements (up to 150 MHz) [29]. Despite those mentioned above, most of the circuit models exclude environmental impacts on the communication channel [1].

**Empirical Models**

Empirical models are then widely exploited to ameliorate the channel’s perception and rectify the shortcomings of the previous techniques. In [30] the path loss of CC-BCC was estimated with the aid of a miniaturized battery-powered transmitter and receiver in addition to exploiting FEM to validate the results. In this work, the correlation between instrument configuration and variations in measurements was specified. For instance, incorporating more ground plane area would undervalue the path loss estimations by up to 33.6 dB. Apart from the instrument configuration, input resistance to the transceivers, cross-sectional area and length of the cables, and variable contact impedances also induce inconsistencies in measurements. Therefore, experiments should be conducted with a high degree of precision to be regarded as reliable [1].
2.2.2 Transceiver Design

The optimized transceiver design in BCC has a pivotal role in facilitating the communication between the transmitter and receiver through the link, i.e., the body. Upon this, there have been multiple attempts targeting the development of BCC transceivers from different design perspectives. The main variations in BCC transceivers are ascribed to the employed coupling method, the operational frequency range, modulation techniques, energy efficiencies, and achievable throughputs [21]. In respect of CC, Zimmerman proposed the first transceiver prototype under PAN [31]. It was implemented using discrete components to verify the viability of communication through the body. The design which supported half-duplex communication was assessed with on-off keying (OOK) and signaling direct sequence spread spectrum (DSSS) modulations for narrow and wide bands, respectively. Despite DSSS’s accuracy, OOK was eventually selected because of its simplicity. As a result, the system achieved a rate of 2.4 kbps for 330 kHz carrier frequency, channel bandwidth 400 kHz, and 1.5 mW power consumption [21]. Later it was proven that the data rates could be extended by adopting other modulation schemes such as frequency-shift keying (FSK) [17].

Today, the state-of-the-art transceivers can achieve high data rates ≥ 1 Mbps while maintaining high energy efficiencies ≤ 1nJ/b. In [13], a CC-BCC transceiver was developed in a 65 nm CMOS process targeting entertainment and healthcare applications. For the entertainment mode, the dual-band (40/160 MHz) full-duplex transceiver can achieve 80 Mbps for an efficiency of 79 pJ/b using binary phase-shift keying (BPSK). While in the healthcare mode, the super-regenerative transceiver employs OOK to attain 100 kbps at 13.56 MHz for low power consumption of 42.5 µW. In [32], a pseudo orthogonal frequency-division multiplexing (P-OFDM) transceiver for hearing aids was implemented in 65 nm CMOS technology. The transceiver utilizes a baseband OFDM that is transmitted through FSK to reduce the effect of interference and signal multipath fading. The transceiver performance displayed a throughput of 1
Mbps with bandwidth 20-120 MHz, low power consumption of 1.4 mW, and 1.4 nJ/bit efficiency. In [33], a broadband transceiver fabricated with 65 nm CMOS technology is proposed. The design is signified with its capability of rejecting interference in the time domain. Further, it proved to satisfy a 30 Mbps throughput and 6.3 pJ/b energy efficiency. Lastly, in this approach, the signal is transmitted as a nonreturn to zero (NRZ) digital data (0/1); hence, the modulation block is eliminated.

### Table 2.1: Comparison of BCC transceivers

<table>
<thead>
<tr>
<th>References</th>
<th>CMOS Process</th>
<th>Frequency Band</th>
<th>Modulation</th>
<th>Rate</th>
<th>Energy/bit</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>HC: 13.56 MHz</td>
<td>OOK</td>
<td>100 kbps</td>
<td>0.43 nJ/b</td>
</tr>
<tr>
<td>[32]</td>
<td>65 nm</td>
<td>20-120 MHz</td>
<td>64 P-OFDM 16 QAM</td>
<td>1Mbps</td>
<td>79 1.4nJ/b</td>
</tr>
<tr>
<td>[33]</td>
<td>65 nm</td>
<td>100 MHz</td>
<td>NRZ</td>
<td>30 Mbps</td>
<td>6.3 pJ/b</td>
</tr>
</tbody>
</table>

### 2.3 Networking

Before tackling networking in CC-BCC-based IoB, we first evaluate the different communication methods in WBAN systems to capitalize on their resource abundance in the literature. The simplest topology to establish communication between nodes is the point-to-point connection. In which a single transmitter conveys information to the receiver over a dedicated link. Hence, the communication is interference resilient [34]. In WBAN, point-to-point communication is widely adopted, where the sensor node measures physiological data and transmits the information to the access point (AP) [35] [36]. However, the use of such a topology is contrary to the objective of IoB. Therefore, multiple access schemes are utilized to enable multipoint communication and mitigate interference. In [37] the authors introduce NOMA under WBAN for the first time to allow two nodes to communicate simultaneously with the hub, i.e., uplink direction. Via packet-level simulations in OMNeT++, it was shown that the NOMA outperforms OMA at high packet rates.
In [38] authors look into access schemes sought to expedite multipoint networks in BCC for the first time. They model OMA and NOMA for a generic IoB network with and without cooperation. Moreover, intending to address QoS demands for different applications, they provide analytical and numerical power control and phase time allocations for three operational regimes: max-min rate, max-sum rate, and QoS sufficient regimes. The numerical results obtained emphasize NOMA’s enhanced performance over OMA in specific scenarios. For instance, at high QoS demands, the transmission power levels in NOMA are less compared to OMA.

Further, in the max-min rate regime, NOMA’s gain over OMA is subject to the maximum transmission power and the relative channel length. In addition, the max-sum rate gain of NOMA increases with transmission power. However, it drops as the gain of the farthest node decrease. Lastly, a comparison between cooperative and non-cooperative schemes in both max-min and max-sum regimes is presented to elicit their performances in various cases. Hence, understanding the network formation is fundamental to support multipoint operation.
Chapter 3

Capacitive Body Channel Access Schemes

We consider an IoB network wherein a wearable hub device (e.g., smartwatch) communicates with $K$ on-body IoB nodes through the time-slotted uplink and downlink transmissions. The smartwatch plays the role of an access point that orchestrates the IoB network and exchanges the information with off-body entities (e.g., smartphones, base stations, routers, etc.) utilizing RF communication methods, e.g., cellular, Bluetooth, Wi-Fi, etc. We should note that the proposed methods are not limited to a specific IoB node deployment over the body, which is a function of the underlying application. Throughout the thesis, we denote the total available UL and DL bandwidth for $K$ IoB nodes, time slot duration, and thermal noise power spectral density by $B$, $T$, and $N_0$, respectively. Additionally, $P_k$ and $P_h$ symbolizes the node’s and the hub’s maximum transmission power, respectively. Without loss of generality, we assume that $P_k$ and $P_h$ are equal to $P$, ensuring the regulatory bodies’ safety compliance. Moreover, the reciprocal channel between $n_k$ and $n_h$ is represented by $g_h^k = g_k^h$.

3.1 Orthogonal Multiple Access (OMA)

The OMA mitigates the multiple access interference by allocating each node with dedicated and equal bandwidths of $B/K$ Hz. In the UL transmission, the received signal from the $k^{th}$ node, $n_k$, at the hub node, $n_h$, on the $k^{th}$ subband is given by

$$y_h^k = \sqrt{P g_h^k g_k^h x_k} + z_h, \ k \in \mathcal{K}$$

(3.1)
where $K$ denotes the index set of all IoB nodes, $\omega^h_k \in [0, 1]$ is the power allocation weight of UL-OMA scheme, $x_k$ is the transmit message of the $k^{th}$ node, $z_h \sim \mathcal{N}(0, N_0B/K)$ represents the additive white Gaussian noise at $n_h$. Accordingly, the signal-to-noise-ratio (SNR) of UL-OMA is given by $\gamma^h_k = \frac{P^h_k \omega^h_k}{N_0B/K}$. Therefore, based on Shannon’s channel capacity theorem the maximum UL achievable throughput for $n_k$ is given by

$$\hat{R}^h_k = \frac{B}{K} \log_2 \left(1 + \gamma^h_k\right), k \in K.$$ (3.2)

Along with dividing the bandwidth, DL OMA divides its power equally among nodes. Which yields $\gamma^h_k = \frac{P^h_k \omega^h_k}{N_0B}$. Accordingly, the DL-OMA scheme can easily be obtained by replacing the terms $(\cdot)_k$ and $(\cdot)^h_k$ with the terms $(\cdot)_h$ and $(\cdot)^h_h$ in (3.1)-(3.2), respectively.

### 3.2 Non-Orthogonal Multiple Access (NOMA)

Unlike OMA, NOMA permits all IoB nodes to transmit their information concurrently over the entire bandwidth $B$. As a result, the UL and DL receiver nodes $n_h$ and $n_k$ perform multi-user detection through successive interference cancellation (SIC), respectively.

#### 3.2.1 Uplink NOMA

In the UL-NOMA, the observed signal at $n_h$ is an aggregation of the transmit signals as follows

$$y_h = \sum_{k \in \tilde{K}} \sqrt{P^h_k \omega^h_k} x_k + z_h,$$ (3.3)

where $\tilde{K}$ denotes the index set of IoB nodes arranged in ascending order of their channel gains, $\omega^h_k \in [0, 1]$ are the UL-NOMA power allocation weights, and $z_h \sim \mathcal{N}(0, N_0B)$. The UL-NOMA scheme allocates power accordingly to ensure that received powers, $P^h_k \triangleq P g^h_k \omega^h_k$, $\forall k$, can be differentiated and follows the same ascending

\[1\]We will consider that all transmit messages satisfy $E(|x_k|^2) = 1$, $\forall k$, throughout the paper.
order, i.e.,

\[
p_1^h < \ldots < p_k^h < \ldots < p_K^h
\]

(3.4)

Accordingly, the SIC receiver decodes and subtracts messages in descending order of their received power. In this way, the \( n_k \) can cancel interference coming from higher rank nodes while the lower rank IoB nodes’ messages are regarded as interference. The SIC receiver may not cancel all the interference due to channel estimation errors and hardware limitations. In this case, the signal-to-interference-plus-noise ratio (SINR) of imperfect SIC and maximum achievable data rate are given by

\[
\zeta_k^h = \frac{P g_k^h \omega_k^h}{\sum_{i=1}^{k-1} P g_i^h \omega_i^h + \epsilon \sum_{j=k+1}^{K} P g_j^h \omega_j^h + N_0 B},
\]

(3.5)

where the first term in the denominator represents the interference resulting from the decoding order’s succeeding messages whereas the second term constitutes the imperfections in SIC, i.e., \( \epsilon \in [0, 1] \) is the residual error coefficient. Accordingly, the achievable UL rate of \( n_k \) is given by

\[
\tilde{R}_k^h = B \log_2 \left(1 + \zeta_k^h\right), k \in \tilde{K}.
\]

(3.6)

### 3.2.2 Downlink NOMA

In the DL-NOMA, \( n_h \) broadcasts a superimposed signal, that is a weighted summation of the intended signals intended for all nodes and received by \( n_k \) as

\[
y_k = \sum_{k \in \hat{K}} \sqrt{P g_k^h \omega_k^h} x_k + z_k,
\]

(3.7)

which is subject to total power consumption constraint, i.e., \( \sum_{k \in \hat{K}} \omega_k^h \leq 1 \) where \( \omega_k^h \in [0, 1] \) are the DL-NOMA power allocation weights and \( \hat{K} \) denotes the index set of IoB nodes arranged in descending order of their channel gains. Notice that the
DL-NOMA follows an opposite decoding order, which yields

the following SINR levels for the SIC receiver

\[ \gamma^k = \frac{P g^k_h \omega^k_h}{\epsilon \sum_{i=1}^{k-1} P g^i_h \omega^i_h + \sum_{j=k+1}^{K} P g^j_h \omega^j_h + N_0 B}, \] (3.8)

Following the SIC procedure, the achievable UL rate of \( n_k \) is given by

\[ \hat{R}^k = B \log_2 \left( 1 + \gamma^k \right), k \in \hat{K}. \] (3.9)

### 3.3 Impacts of Decoding Order on Energy Efficiency, Fairness, and Network Lifetime

In [39, 40, 41], a great emphasis was put on the effect of decoding-order to optimize the performance of the NOMA scheme over OMA. As it was proven that descending and ascending channel gain ordering provides a significant gain. Noting that the objective was to deliver maximum sum-throughput; hence, we reverse the ordering to achieve lower power consumption and higher fairness, which mainly determines the IoB network lifetime. To quantify these key performance metrics, let us first model the power consumption of IoB nodes as follows

\[ P^{ul}_{k,c} = P^{crc}_{k} + P^{tx}_{k} = E^{crc}_{k} R^{h}_{k} + P^{\omega}_{h}, \] (3.10)

where \( P^{crc}_{k} \) and \( P^{tx}_{k} \) are the power consumed for circuit and transmission. In the BCC transceiver design literature, the energy efficiency of transceivers is often measured by the energy consumed per transmitted bit. Therefore, \( P^{crc}_{k} \) can be defined as multiplication of transceiver efficiency, \( E^{crc}_{k} [\text{Joules/bit}] \), and data rate \( R^{h}_{k} [\text{bps}] \). On the other hand, effective transmission power is simply weighted maximum transmission power, \( P^{\omega}_{h} \), where we omit (\( \square \)) and (\( \square \)) notations to capture both UL-OMA and
UL-NOMA schemes. Similarly, the power consumption of the hub is given by

$$P_c^h = \sum_{k=1}^{K} P_{k,c}^{dl} = \sum_{k=1}^{K} E_{h,k}^{crc} R_{h,k} + P_{h,k}.$$  \hspace{1cm} (3.11)

Fairness metrics have a pivotal role in assessing the performance of the wireless systems in terms of resource sharing. Accordingly, we exploit Jain’s index \[42\] to measure the UL power consumption fairness as follows:

$$J(P_{ul,1,c}, P_{ul,2,c}, \ldots, P_{ul,K,c}) = \frac{\left(\sum_{k=1}^{K} P_{ul,k,c}\right)^2}{K \sum_{k=1}^{K} (P_{ul,k,c})^2} \in [0, 1],$$ \hspace{1cm} (3.12)

which can be rewritten for the DL case by replacing $P_{k,c}^{dl}$ with $P_{k,c}^{ul}$. Assuming that identical IoB nodes, we define the network lifetime metric as the time span between network initialization and the time slot when the first battery depletion occurred, i.e.,

$$NL = \min_{\forall k} \left\{ \frac{B_{k}^{int}}{\alpha_k P_{k,c}^{ul} T} \right\},$$ \hspace{1cm} (3.13)

where $B_{k}^{int}$ is the initial battery level, and $\alpha_k$ is the transmission duty cycle of $n_k$. In Chapter \[4\] we will provide a thorough numerical analysis of these performance metrics.

### 3.4 Problem Formulation and Solution Methodology

#### 3.4.1 Problem Formulation

It is obvious from (3.13) the network lifetime is mainly determined by power consumptions given in (3.10) and (3.11). The optimization problem that optimizes the power allocation weights to minimize total UL power consumption can be formulated
as

\[ P_{UL} : \min_{0 \leq \omega \leq 1} \sum_{k \in K} P^k_c, \]  
\[ C_1: \quad \text{s.t.} \quad R^h_k(\omega) \geq R^h_k, \quad \forall k \]  

where \( C_1 \) is the QoS constraints that ensure that \( n_k \) is provided with a data rate not less than its demand \( R^h_k \) and \( \preceq \) denotes the pairwise inequality. Similarly, the DL problem can be formulated as

\[ P_{DL} : \min_{0 \leq \omega \leq 1} P^h_c \]  
\[ C_1: \quad \text{s.t.} \quad R^h_k(\omega) \geq R^h_k, \quad \forall k, \]  
\[ C_2: \quad \sum_k \omega^k_h \leq 1 \]

where \( C_2 \) is an additional constraint to ensure total DL transmission power is less than the maximum transmission power of \( n_h \). Both (3.14) and (3.15) can be readily solved put into convex optimization solvers. However, considering the low-cost and ultralow-power design goals of IoB nodes, it becomes essential to derive closed-form optimal power allocations to reduce hardware cost and power consumption related to the computational complexity.

### 3.4.2 Solution Methodology

Both \( P_{UL} \) and \( P_{DL} \) reach an optimal point when QoS constraints are active, i.e., satisfied with equality, because providing a data rate more than demanded increases both circuit and transmission power consumption. Then, these problems can be modeled as follows \[ 43 \]

\[ (I - \Gamma J) \mathbf{p} = \bar{\Gamma} \mathbf{\sigma}, \]  

where \( I, \bar{\Gamma} = \text{diag}(\bar{\Gamma}_1, \ldots, \bar{\Gamma}_k, \ldots, \bar{\Gamma}_K) \) and \( J \) denotes the identity matrix, diagonal matrix of the SINR demands consistent with QoS demands, and interference channel
gain matrix, respectively. Similarly, \( \mathbf{p} \) and \( \mathbf{\sigma} \) refer to column vectors of the received powers and receiver noise, respectively. The vectors are of dimension \( K \times 1 \), and matrices are \( K \times K \) dimensions. Moreover, the entries of \( \mathbf{J} \) take the following definition

\[
\begin{aligned}
J_{ij}^H &= \begin{cases}
0, & i < j \\
0, & i = j \\
0, & i > j \\
\epsilon, & i < j \\
1, & i > j \\
1, & i < j \\
0, & i = j \\
\epsilon, & i > j
\end{cases} \\
&= \begin{cases}
\text{OMA} & \text{if } i < j, j < i \\
\text{UL-NOMA} & \text{if } i = j, j = i \\
\text{DL-NOMA} & \text{if } i > j, j > i
\end{cases}
\end{aligned}
\] (3.17)

The entries 0, 1, and \( \epsilon \) indicates no interference, cluster-interference, and residual interference, respectively \[44\]. For UL direction (3.16) is contingent to \( p_k \leq P \sum_k g_{hk}^h \) due to \( \omega_{hk}^h \leq 1 \). Likewise, in the DL, (3.16) is also subject to \( \sum_k p_k \leq P \sum_k g_{hk}^k \) due to \( \sum_k \omega_{hk}^k \leq 1 \) plus \( p_k \leq P \sum_k g_{hk}^k \) due to \( \omega_{hk}^k \leq 1 \).

According to \textit{Perron-Frobenius theorem}, describing non-negative irreducible matrices, the eigenvalue of \( \mathbf{J} \) is then real and positive. Additionally, the eigenvector corresponding to it is also non-negative \[45\]. To attain feasible solution for (3.16), the maximum eigenvalue of \( \mathbf{H} \triangleq \bar{\Gamma}(\theta) \mathbf{J} \) is restricted to be less than unity \[46\]. Assuming feasible QoS demands, the solution for (3.16) is then given by \( \mathbf{p}^* = (\mathbf{I} - \Gamma \mathbf{J})^{-1} \bar{\Gamma} \mathbf{\sigma} \), from which the optimal power allocation weights can be obtained as follows:
OMA

Defining the SINR constraint by $\gamma^h_k \equiv 2\frac{\bar{\rho}^h_k}{P_k} - 1$, the optimal power allocations for UL-OMA and DL-OMA are given by

$$
\omega^h_k,\star = \frac{\tilde{\gamma}^h_k N_0 B}{g^h_k K P_k}, \forall k \in \mathcal{K}, \text{ and}
$$

$$
\omega^k,\star = \frac{\tilde{\gamma}^h_k N_0 B}{g^h_k P}, \forall k \in \mathcal{K},
$$

which are subject to $\omega^h_k,\star \leq 1, \forall k$, and $\omega^k,\star \leq 1, \forall k$, respectively.

UL-NOMA

Defining the SINR constraint by $\gamma^h_k \equiv 2\frac{\bar{\rho}^h_k}{P_k} - 1$, the optimal power allocation for the UL-NOMA is given by

$$
\omega^{h,\star} = \frac{\tilde{\gamma}^h_k N_0 B}{g^h_k P g^h_k} \left(1 - \epsilon \frac{\gamma^h_k}{1 + \gamma^h_k}\right)^{K-k} \left(1 + \epsilon \frac{\gamma^h_k}{1 + \gamma^h_k}\right)^{K-k}, \forall k \in \mathcal{K},
$$

which can be simplified for the perfect case ($\epsilon \to 0$) as follows

$$
\omega^{h,\star} = \frac{N_0 B}{P g^h_k} \gamma^h_k (1 + \gamma^h_k)^{k-1}, \forall k \in \mathcal{K}.
$$

DL-NOMA

Defining the SINR constraint by $\gamma^k_h \equiv 2\frac{\bar{\rho}^k_h}{P_k} - 1$, the optimal power allocation for the UL-NOMA is given by

$$
\omega^{1,\star} = \frac{\tilde{\gamma}^1_h N_0 B}{g^1_h P g^1_h} \left(1 - \epsilon \frac{\gamma^1_h}{1 + \gamma^1_h}\right)^{K-1} \left(1 + \epsilon \frac{\gamma^1_h}{1 + \gamma^1_h}\right)^{K-1}
$$

Equations (3.20)-(3.24) are obtained following the derivation in [47].
\[
\hat{\omega}^k_\star = \frac{\hat{\omega}^{1,\star}_h}{P g_h^k (1 + \gamma_h^k)^{k-1}}, \forall k \in \hat{K}.
\] (3.23)

which can be simplified for the perfect case \((\epsilon \to 0)\) as follows

\[
\hat{\omega}^k_\star = \frac{N_0 B}{P g_h^k \gamma_h^k (1 + \gamma_h^k)^{K-k}}, \forall k \in \hat{K}.
\] (3.24)

### 3.5 Energy Harvesting

In conventional IoB networks, batteries are the primary energy source to drive both sensing and transmission processes. However, due to the battery capacity limitations, the network lifetime’s adequacy is then significant. On that account, there have been multiple proposed solutions to impede battery depletion. Primarily, duty cycle-based operation presented in (3.13), where the nodes are periodically put into sleep mode, to reduce energy consumption in the network. Although adopting such protocol proved to be successful in increasing the longevity of batteries yet, operating at low duty cycles does not aid the network’s performance. That is, low duty cycles correspond to an increase in transmission latency and decrease of the throughput. To address energy scarcity, many researchers in wireless sensor networks are deploying energy harvesters in their systems. In this context, we will examine IoB nodes equipped with RF energy harvesting rectennas. The energy harvested from source \(m\) is modeled as

\[
P_{\text{harv}}^m = P^m_x g_x^m \eta
\] (3.25)

where \(P^m_x\) is the signal strength in dBm, \(g_x^m\) is the linear channel gain obtained from the free space path loss [48], and \(\eta \in [0, 1]\) is the power conversion efficiency (PCE). In our simulations, we will consider both single-band and multi-band rectenna. Thus,
for multi-band rectenna, the total harvested energy is defined as

\[ P_{\text{harv}}^{\text{tot}} = \sum_{m \in \mathcal{M}} \beta_m P_{\text{harv}}^m \]  

(3.26)

hence, \( \mathcal{M} \) denotes the set of \( m \) RF sources. To indicate the energy arrival rate, we introduce the duty cycle \( \beta_m \in [0, 1] \). The battery level evolution of the entire duration for UL direction is depicted in the following equation

\[ B^k(t + 1) = \min \{ B_k^{\text{max}}[k], [B^k(t) + T (P_{k,c}^{\text{ul}} - P_{\text{harv}}^{\text{tot}})] \}^+ \]  

(3.27)

Towards obtaining reliable results, we govern the battery level by imposing the minimum argument. Further, to obtain the DL battery level evolution we simply replace \( P_{k,c}^{\text{ul}} \) with \( P_{k,c}^{\text{dl}} \) both are defined in (3.10).

### 3.6 Maximum Feasible Number of Nodes

In this section, we obtain the maximum number of nodes, \( K_{\text{max}} \), that can be admitted in the IoB system given the total power constraint, throughput requirements, and bandwidth. This is essential to maintain the validity of the power control methods.

#### 3.6.1 OMA

For UL-OMA, the violation of power control occurs if at least one node is demanding a weight that exceeds unity as per (3.18). Since the lowest channel gain node is assigned the highest weight, then \( K_{\text{max}} \), in this case, is derived from the inequality

\[ \bar{\gamma} N_0 B \leq 1 \]

Where, \( \bar{\gamma} \) symbolizes the lowest channel gain in the network. In consideration of analytical tractability, we assume that \( \gamma^h_k \gg 1 \) and approximate

\[ \gamma^h_k \triangleq 2^{\frac{\beta_k K}{B}} - 1 \approx \left[ x \right]^+ = \max(0, 1) \]
\(2^{-\frac{\bar{R}_k}{\bar{B}}}\). Solving the inequality, then, yields

\[
K_{\text{max}} = \left\lfloor \frac{W_{-1} \left( -\frac{NaR \log(2)}{gP_\bar{R}} \right) B}{\log(2)R} \right\rfloor,
\]

(3.28)

where \(W_{-1}(\cdot)\) is the \(-1\)th branch of Lambert-W function. For the DL-OMA, \(K_{\text{max}}\) can be obtained directly from equation (3.2) as

\[
K_{\text{max}} = \left\lfloor \frac{B}{R} \log_2 \left( 1 + \frac{P_\bar{g}}{N_0B} \right) \right\rfloor.
\]

(3.29)

### 3.6.2 UL-NOMA

In NOMA schemes for both directions we consider perfect SIC and identical QoS demands, \(\bar{R}\), hence, identical SINR thresholds \(\bar{\gamma}\). According to (3.21), the highest channel gain node will be the first node to breach power constraints. Thus, \(K_{\text{max}}\) is found from \(\frac{NaR}{P_\bar{g}} \bar{\gamma}(\bar{\gamma} + 1)^{k-1} \leq 1\) as follows

\[
K_{\text{max}} = \left\lfloor 1 + \frac{\log \left( \frac{P_\bar{g}}{\bar{\gamma}N_0B} \right)}{\log(1 + \bar{\gamma})} \right\rfloor,
\]

(3.30)

where \(\bar{g}\) is the maximum channel gain in the IoB network.

### 3.6.3 DL-NOMA

Conversely, the maximum number of nodes for DL-NOMA \(K_{\text{max}}\), is obtained from the constraint imposed on the total power \(\sum_{k=1}^{K} \omega_k \leq 1\). As per (3.24), the highest channel gain node is correlated with the highest power coefficient, which contributes the most to the sum of weights. By assuming other nodes has the same highest
channel gain, \( \bar{g} \), the total weight can be approximated as follows:

\[
\sum_{k=1}^{K} \omega_k = \sum_{k=1}^{K} \rho \bar{\gamma} (\bar{\gamma} + 1)^K \left( \frac{1}{(\bar{\gamma} + 1)^k} \right) = \sum_{k=1}^{K} a \varrho^k
\]  

(3.31)

where \( \varrho \equiv \frac{1}{(\bar{\gamma} + 1)} \) and \( a = \rho \bar{\gamma} \varrho^{-K} \). By setting \( m = 1 \) and \( n = K \) in the geometric progression formula, i.e., \( \sum_{k=m}^{n} a \varrho^k \equiv a \frac{(\varrho^n - \varrho^{n+1})}{1 - \varrho} \), (3.31) can be rewritten as

\[
\sum_{k=1}^{K} \omega_k = \frac{\rho \bar{\gamma} \varrho^{-K} (\varrho - \varrho^{K+1})}{1 - \varrho} = \frac{\rho \bar{\gamma}}{1 - \varrho} (\varrho^{1-K} - \varrho)
\]  

(3.32)

By substituting \( \varrho \equiv \frac{1}{(\bar{\gamma} + 1)} \) into (3.32), we obtain \( \sum_{k=1}^{K} \omega_k = \rho \left[(1 + \bar{\gamma})^K - 1\right] \leq 1 \), which yields

\[
K_{\text{max}} = \left\lfloor \frac{\log(\frac{\rho + 1}{\rho})}{\log(1 + \bar{\gamma})} \right\rfloor.
\]  

(3.33)
Chapter 4

Cooperative Capacitive Body Channel Access Schemes

4.1 Cooperative Orthogonal Multiple Access

4.1.1 Uplink C-OMA

As presented in [38], the cooperative OMA is executed in two phases where the hub communicates to source nodes through the relay. By which, the highest channel gain node \( n_{\text{max}} \) is designated as the relay. In the preceding phase, the relay is switched to idle status to acquire and decode the signals of \( K - 1 \) source nodes over \( \lambda T \) duration. Thus, the received signal by the relay is

\[
y_n^k = \sqrt{P g_k^n \omega_n^k} x_k + z_n, \quad k \in \mathcal{K}_{n-1}
\]  

(4.1)

where \( \mathcal{K}_{n-1} \) denotes the index set of source nodes, \( \omega_k^n \in [0, 1] \) is the power allocations of UL-OMA scheme in the first phase, \( z_n \sim \mathcal{N}(0, N_0B/(K-1)) \). Then the SNR and data rate correlated to this phase are

\[
\hat{R}_n^k = \frac{B}{K - 1} \log_2 \left( 1 + \hat{\gamma}_n^k \right), \quad k \in \mathcal{K}_{n-1}.
\]  

(4.2)

\[
\hat{\gamma}_n^k = \frac{P g_k^n \omega_k^m}{N_0B/(K-1)}
\]  

(4.3)

In the later phase, the status of the IoB nodes is reversed. Hence, the relay is on to transmit the received messages along with its own to the hub node in \( (1 - \lambda)T \)
time length. Thence, the observed signal at the hub is given by

\[ y_{k,n}^h = \sqrt{P g_{h_n}^h \hat{\omega}_{k,n}^h} \hat{x}_k + z_h, \; k \in \mathcal{K} \]  \hspace{1cm} (4.4)

where \( \hat{\omega}_{k,n}^h \in [0, 1] \) is the power control level assigned by \( n_n \) to communicate \( n_k \)'s message, \( \hat{x}_k \) is the decoded message of the \( n_k \). In this case, the data rate of this phase is

\[ \hat{R}_{k,n}^h = \frac{B}{K} \log_2 \left( 1 + \hat{\gamma}_{k,n}^h \right), \; k \in \mathcal{K}. \]  \hspace{1cm} (4.5)

where the SNR is defined as \( \hat{\gamma}_{k,n}^h = \frac{P g_{h_n}^h \hat{\omega}_{k,n}^h}{N_0 B/K} \). The end-to-end data rate of this relayed link is

\[ \hat{R}_k^h(\hat{\lambda}) = \min \left( \hat{\lambda} \hat{R}_n^h, (1 - \hat{\lambda}) \hat{R}_{k,n}^h \right), \; \forall k \in \mathcal{K} \]  \hspace{1cm} (4.6)

4.1.2 Downlink C-OMA

In the DL C-OMA case, the maximum transmission power of \( n_h \) is equally split between nodes. Thereby the SNR of the transmitted message from \( n_h \) to \( n_n \) is expressed as \( \hat{\gamma}_{h_n}^{n,k} = \frac{P g_{h_n}^n \hat{\omega}_{n,k}^{n,k}}{N_0 B} \). Where \( \hat{\omega}_{h_n}^{n,k} \) is the power allocated by \( n_h \) to transmit \( n_k \)'s message through \( n_n \). Further, DL-COMA scheme of the first phase can easily be obtained by replacing the term \( (\cdot)^{h,n}_{k,n} \) and \( (\cdot)_z \) with \( (\cdot)^{n,k}_{h_n} \) and \( (\cdot)^{n}_{n} \) in (4.4)-(4.5), respectively. Similarly, in phase two \( n_n \)'s power is divided between source nodes, thusly, the SNR of the communicated signal is \( \hat{\gamma}_{n}^k = \frac{P g_{n_h}^k \hat{\omega}_{n,k}^n}{N_0 B} \). In a similar manner, the DL scheme in the second phase is found by substituting \( (\cdot)^{k}_{n} \) and \( (\cdot)_k \) in place of \( (\cdot)^{n}_{n} \) and \( (\cdot)^{k}_{k} \) in equations (4.1)-(4.2). It is important to note that the end-to-end rate also follows equation (4.6).
4.2 Cooperative Non-Orthogonal Multiple Access

4.2.1 Uplink C-NOMA

In UL cooperative NOMA, the time slot is divided into two phases. The first stage is formed by the relay node, which is the strongest user in the system, i.e., the highest channel gain user and $K - 1$ source nodes. In which the relay node $n_n$ is put in idle mode to listen to transmissions from other nodes over $\lambda T$ time duration. Accordingly, $n_n$ performs SIC to retrieve each message superimposed by other signals. In this case, the SINR and throughput in the first phase are given by

$$\gamma^n_{k} = \frac{P g_k^n \hat{\omega}^n_k}{\sum_{i=1}^{K-1} P g_i^n \hat{\omega}^n_i + \epsilon \sum_{j=k+1}^{K} P g_j^n \hat{\omega}^n_j + N_0 B},$$

$$\hat{R}^n_k = B \log_2 (1 + \gamma^n_k), k \in \hat{K}_n.$$ (4.7)

where $\hat{K}_n$ indicates the index set of source nodes arranged in ascending order of their channel gains, $\hat{\omega}_k^n \in [0, 1]$ are the UL-NOMA power control levels of the first phase.

In the second phase, the source nodes are switched to idle whereas the relay node is activated to broadcast its message along with the recovered messages to the hub node $n_h$ in the reminder time slot, $(1 - \lambda)T$. Respectively, the received signal at the hub is expressed as

$$y_h = \sum_{k \in \hat{K}} \sqrt{P g_h^n \hat{\omega}^h_{k,n}} \hat{x}_k + z_h,$$ (4.9)

where $z_h \sim \mathcal{N}(0, N_0 B)$. To extract each signal, the hub node $n_h$ also performs SIC, which yields an SINR and data rate of

$$\gamma^n_{k,n} = \frac{P g^n_h \hat{\omega}^h_{k,n}}{\sum_{i \in \hat{K}} P g_i^n \hat{\omega}^n_i + \epsilon \sum_{j=k+1}^{K} P g_j^n \hat{\omega}^n_j + N_0 B},$$ (4.10)
\[ \hat{R}_{k,n}^h = B \log_2 \left( 1 + \hat{\gamma}_{k,n}^h \right), \quad k \in \hat{\mathcal{K}}. \]  

(4.11)

where \( \hat{\mathcal{K}} \) is the index set of all IoB nodes in the network in ascending sequence of their channel gains, \( \hat{\omega}_{k,n}^h \in [0, 1] \) are the UL-NOMA power allocations of the second phase. It is worth noting that for this two-phase network, the end-to-end data rate for \( n_k \) is

\[ \hat{R}_k^n(\check{\lambda}) = \min \left( \check{\lambda} \hat{R}_k^n, (1 - \check{\lambda}) \hat{R}_{k,n}^h \right), \quad \forall k \in \hat{\mathcal{K}} \]  

(4.12)

that is, it is the transmission throughput of the bottleneck link.

### 4.2.2 Downlink C-NOMA

Similarly, the DL cooperative NOMA is attained in two phases. The first phase is the direct transmission phase, where the hub node broadcasts the superposed messages to the relay node that is put in idle for \( \lambda T \) duration. The received signal by \( n_n \) is

\[ y_n = \sum_{k \in \mathcal{K}} \sqrt{P g_n^h \hat{\omega}_n^k} x_k + z_n, \]  

(4.13)

where the total power consumption of the first phase is governed by the constraint, i.e., \( \sum_{k \in \hat{\mathcal{K}}} \hat{\omega}_n^k \leq 1 \). Hence, \( \hat{\omega}_n^k \in [0, 1] \), and \( \hat{\mathcal{K}} \) is the index set of IoB nodes arranged in descending order of their channel gains. Therefore, the SINR and data rate in this phase following SIC implementation by \( n_n \) are

\[ \hat{\gamma}_n^k = \frac{P g_n^h \hat{\omega}_n^k}{\epsilon \sum_{i=1}^{k-1} P g_{h_i}^n \hat{\omega}_{n,i} + \sum_{j=k+1}^{\hat{\mathcal{K}}} P g_{h_j}^n \hat{\omega}_{n,j} + N_0 B}, \]  

(4.14)

\[ \hat{R}_n^h(k) = B \log_2 \left( 1 + \hat{\gamma}_n^k \right), \quad k \in \hat{\mathcal{K}}. \]  

(4.15)
Moreover, in the second phase, \( n_n \) relays the aggregated message to source nodes over \((1 - \lambda)T\) time period. Thus, each source node withdraws its signal by performing SIC. As a result, the SINR levels for the SIC receiver and the corresponding throughput are

\[
\gamma^{k}_{n} = \frac{P_{g_{n}}^{k}g_{n}^{k}}{\sum_{i=1}^{k-1} P_{g_{n}}^{i}g_{n}^{i} + \epsilon \sum_{j=k+1}^{K} P_{g_{n}}^{k}g_{n}^{k} + N_{0}B}, \quad (4.16)
\]

\[
\tilde{R}_{n}^{k} = B \log_{2} (1 + \gamma^{n}_{k}), \quad k \in \hat{K}_{-n}. \quad (4.17)
\]

Additionally, the data rate of the link is subject to equation (4.12).

### 4.3 Problem Formulation and Solution Methodology

#### 4.3.1 Problem Formulation

In contrast to regular access schemes, optimized power and time allocations are requisite for the success of cooperative schemes. Hence, the objective function, i.e., the total consumed energy in both phases, is found

\[
E_{UL} : \min_{\lambda} \quad \lambda T \sum_{k \in K_{-n}} \omega_{n}^{k} + (1 - \lambda) T \sum_{k \in K} \omega_{k,n}^{h}
\]

\[
C_{1} : \quad \text{s.t.} \quad R_{k,n}(\omega) \geq \tilde{R}_{k}^{h}/\lambda, \quad \forall k, \quad \text{and} \quad (4.18)
\]

\[
C_{2} : \quad R_{k,n}(\omega) \geq \tilde{R}_{k}^{h}/(1 - \lambda), \quad \forall k
\]

Where \( C_{1} \) and \( C_{2} \) are the QoS constraints in the first and second phases that satisfy the bottleneck link throughput. We also exclude (\( \hat{\square} \)) and (\( \tilde{\square} \)) to realize both
OMA and NOMA schemes. Likewise, the DL problem is formulated as

\[
E_{DL} : \min_{\lambda} \lambda T \sum_{k \in \mathcal{K}} \omega_{n,k}^h + (1 - \lambda) T \sum_{k \in \mathcal{K} \cup \mathcal{N}} \omega_{n,k}^h
\]

\[C_1: \mbox{s.t.} \quad R_{t,h}^{n,k}(\omega) \geq \bar{R}_{t,h}^h/\lambda, \ \forall k\]

\[C_2: \quad R_{n}^{h}(\omega) \geq \bar{R}_{h}^{k}/(1 - \lambda), \ \forall k\]

\[C_3: \quad \sum_{k} \omega_{n,k}^h \leq 1\]

\[C_4: \quad \sum_{k} \omega_{n,k}^h \leq 1\]

where \(C_3\) and \(C_4\) are additional constraints to guarantee that the total DL transmission power in both stages is within the maximum transmission power limit.

### 4.3.2 Solution Methodology

In order to solve the joint phase time and power optimization problem, we activate all of our constraints with equality, i.e., satisfying exact QoS demands prolongs the battery lifespan. This is applicable to UL and DL transmissions in OMA and NOMA schemes. The solution is expeditiously obtained using Golden Section search to update the value of \(\lambda\) iteratively. In each run, \(\lambda\) value is used to calculate the rates in the first and second phases of the transmission link from \(C_1\) and \(C_2\), respectively. The optimal power weights are, then, attained by substituting the rates in equations (3.18) - (3.20) for UL-OMA and UL-NOMA or (3.19) - (3.23) for DL-OMA and DL-NOMA. The process is repeated until finding the minimum, which satisfies (4.18). The algorithm structure to find optimal \(\lambda\) on the interval \([a, b]\) is presented in Algorithm 1. Where \(\tau\) and \(N\) denote the Golden Ratio and maximum iterations, respectively.
Algorithm 1: Overall algorithm for optimal \( \lambda \)

Initialize \( a, b, \tau, N, \text{Tolerance} \);

Compute \( \lambda \) as:

\[
\lambda_1 = a(1 - \tau) + b\tau, \quad \lambda_2 = a\tau + b(1 - \tau);
\]

\textbf{while} \( |b - a| > \text{Tolerance and } i < N \) \textbf{do}

Implement, Algorithm 2 to find:

\( f(\lambda_1) \) and \( f(\lambda_2) \);

Update \( \lambda \) using Golden-Section;

\( i++ \);

\textbf{end}

To compute energy consumption for UL-OMA and UL-NOMA the following algorithm 2 is used. To extend it for DL, \( (\cdot)^{k}_{n} \) and \( (\cdot)^{h}_{k,n} \) are replaced with \( (\cdot)^{n,k}_{h} \) and \( (\cdot)^{h}_{n} \).

Algorithm 2: Total energy consumption in NOMA and OMA algorithm

Initialize \( \bar{R}^{h}_{k}, P, B, K, \text{gain vector } G \);

Compute \( R^{n}_{k} \) and \( R^{h}_{k,n} \) from \( C_1 \) and \( C_2 \) in (4.18);

Evaluate \( \gamma^{n}_{k} \) and \( \gamma^{h}_{k,n} \);

OMA from (4.2) and (4.5), NOMA from (4.8) and (4.11);

Obtain \( \omega^{n}_{k} \) and \( \omega^{h}_{k,n} \);

OMA from (3.18) and (3.19), NOMA from (3.20) and (3.23);

Solve \( f(\lambda) = \lambda T \sum_{k \in K_{n}} \omega^{n}_{k} + (1 - \lambda) T \sum_{k \in K} \omega^{h}_{k,n} \);
Chapter 5

Results and Discussion

In this chapter, simulation results are illustrated to assess the performance of the proposed energy-efficient body channel access schemes as we investigate various network design parameters. Unless stated otherwise, parameters available in Table 5.1 will be exploited throughout the simulations.

Table 5.1: Simulation Parameters

<table>
<thead>
<tr>
<th>Par.</th>
<th>Value</th>
<th>Par.</th>
<th>Value</th>
<th>Par.</th>
<th>Val.</th>
<th>Par.</th>
<th>Val.</th>
</tr>
</thead>
<tbody>
<tr>
<td>( B )</td>
<td>1 MHz</td>
<td>( N_0 )</td>
<td>-174 dBm/Hz</td>
<td>( B_k^{\text{init}} )</td>
<td>1 Joules</td>
<td>( K )</td>
<td>3</td>
</tr>
<tr>
<td>( T )</td>
<td>1 sec.</td>
<td>( P_m )</td>
<td>-30 dBm</td>
<td>( \bar{R} )</td>
<td>1 Mbps</td>
<td>( \epsilon )</td>
<td>( 1e^{-6} )</td>
</tr>
</tbody>
</table>

5.1 Non-Cooperative Access Schemes

5.1.1 Closed-form and Numerical solver

From the proceeding discussion in chapter 3, we learn the vital role of power allocation in NOMA, hence, the foremost objective is to examine the presented power allocation algorithm. In this regard, we begin with comparing the sum of power weights obtained via the proposed CF solution and CVX. Here we consider a network that can support 10 IoB nodes all having equal QoS demands. The nodes are deployed such that the nearest node to the hub is \( \ell_1^h = 10 \) cm away, the \( k^{th} \) node is then \( k \times 10 \) cm away from \( n_h \). It is clear from Fig. 5.1, the results attained by both approaches match tightly displaying the positive association between high power consumption, high QoS demands, and larger network size. Noting that the presented results are for UL traffic.
but a similar behaviour is observed for DL.

Figure 5.1: Sum of power weights obtained by proposed CF solution and CVX w.r.t QoS and $K$

Further, since sensor nodes are operating on a limited power budget, than a fair power control is important to prolong the network life time. Therefore, we compare the fairness metric for both CF and CVX as a function of cancellation error $\epsilon$ and varying channel conditions between the nodes (i.e., their locations). Which is modelled as the ratio between the first and third nodes’ channel lengths $\ell_3^h/\ell_1^h$. We set the channel lengths for the $\ell_1^h$ and $\ell_2^h$ at 10 and 20 cm respectively, while we sweep $\ell_3^h$ from 30 to 180 cm. As depicted by Fig. 5.2 CF performs better than CVX in terms of fair power allocation between IoB nodes. In both cases the fairness reduces as we push $n_3$ further away from the hub. Interestingly, the fairness metric reaches the maximum value in both CF and CVX as $\epsilon \to 1$ because $\epsilon = 1$ yields an interference OMA network, where all nodes transmit at maximum power without power control. Nonetheless, this fairness does not necessarily imply energy efficiency as $\epsilon = 1$ also yields the higher power consumption. Moreover, the impact of channel length ratio $\ell_3^h/\ell_1^h$ at $\epsilon = 1$ reduces fairness drastically for CF unlike CVX. Lastly,
adopting the CF power control reduces the processing complexity reflected in lower power consumptions.

![Figure 5.2: Fairness comparison between proposed CF solutions and CVX w.r.t SIC and channel gain ratio](image)

**5.1.2 Power Allocation Factors NOMA and OMA**

Figure 5.3 plots the power allocation factors for OMA in comparison with NOMA for UL and DL transmissions in 5.3a and 5.3b respectively. For the UL case, it is shown that weights for both schemes are closely matched for low QoS demands. As the demands and network size increase, the weights in NOMA schemes are somewhat less than the OMA scheme. In contrast, for DL, NOMA significantly reduces power consumption. This gain is due to the fact that in DL, the maximum transmission power of $n_h$ is equally distributed between nodes. Hence, the superiority of NOMA over OMA is more pronounced.
Figure 5.3: Sum Power Weights w.r.t QoS and $K$: a) UL-OMA vs. UL-NOMA and b) DL-OMA vs. DL-NOMA.

5.1.3 Network Lifetime

Proceeding with CF optimal power allocation and based on an initial % 100 battery capacity, we present the network lifetime in Fig. 5.4. The network life time is attained for identical nodes with respect to their transmission duty cycle and QoS demands. The nodes are equipped with 1 J battery deriving both the circuitry and data transmission. Since a considerable amount of energy is consumed by the circuit, we look at two different transceiver energy efficiencies. In both cases, the results agree with the intuition that lower $\alpha$ values (i.e., lower energy departure rate) and lower QoS demand results in a longer lifespan. Moreover, deploying a transceiver with 1 pJ/bit energy efficiency will increase the life span by $10^3$ slots compared to 1 nJ/bit. That is, the overall node lifetime of IoB devices is primarily determined by the circuit’s energy efficiency rather than transmission power. Hence, the significance of circuit design optimization.
5.1.4 Maximum Feasible Number of Nodes

In Fig. 5.5 and Fig. 5.6, we determine the maximum number of IoB nodes that can be accommodated in an IoB network operating on OMA and NOMA access schemes against QoS and maximum transmission power constraints, respectively. Further, in both figures, we illustrate both analytical and simulation results for $K_{\text{max}}$. Throughout the analysis, we investigate two channel lengths assigned to the highest channel gain node $\ell_{h1} = 20$ cm and $\ell_{h1} = 100$ cm, while maintaining the lowest channel gain node at $\ell_{hK} = 180$ cm for all cases. The subsequent nodes will follow a uniform distribution as $\ell_{hK} = \ell_{h1} + 5(K - 1)$. From Fig. 5.6a, we note an inverse relationship between the number of nodes and QoS requirements for both OMA and NOMA.

Yet, the most important observation is that indeed the number of nodes in NOMA is dependent on $\ell_{h1}$. On the contrary, OMA is unaffected by such variations because $K_{\text{max}}$ is governed by $n_{K}$ at 180 cm. Further, the eminence of NOMA over OMA from the network capacity perspective is also evident. Fig. 5.6b provides maximum nodes analysis for DL direction displaying a similar behaviour to results in 5.6a.
Interestingly, the OMA case showed a degree of sensitivity to $\ell_1^h$.

![Graphs showing $K_{\text{max}}$ vs. QoS for UL-OMA vs. UL-NOMA and DL-OMA vs. DL-NOMA.](image)

Figure 5.5: $K_{\text{max}}$ vs. QoS: a) UL-OMA vs. UL-NOMA and b) DL-OMA vs. DL-NOMA.

Fig. 5.6 provides the analysis of the number of nodes in UL transmission subject to maximum transmission powers from -40 dBm to 0 dBm. It is obvious that the number of nodes increases as the power increases.
Figure 5.6: $K_{\text{max}}$ vs. QoS: a) UL-OMA vs. UL-NOMA and b) DL-OMA vs. DL-NOMA.

5.1.5 Battery Level Evolution

Figure 5.7 plots the battery capacity evolution with time for a node equipped with harvesting circuitry employing two different transceiver efficiencies. Additionally, the rectenna with 25% PCE can behave both as a single-band or multi-band harvester. The figure shows charging trends for $\alpha = 0.5$ and $\beta = 0.5$ for an initial battery capacity of 10% for the different operational modes. For the first energy consumption, 1PJ (solid lines), only devices close to the node could provide sufficient energy. Whereby the battery can profit from this and store the excess energy. It further illustrates that multi-band energy harvesting can outshine single-band harvesting. Nevertheless, none of the sources could surpass the second power consumption 0.5 nJ (dashed lines).
5.2 Cooperative Access Schemes

For cooperative schemes’ simulations we consider a network formed by source nodes, a relay $n_n$ and the hub node $n_h$. We assume all nodes have the same QoS requirements of 0.5 Mbps yet they vary in their channel gains. The distance between hub and relay, relay and source nodes, and number of nodes in the system vary in simulation so as to study the schemes performance for different scenarios. The results in Figures 5.8 - 5.9 indicate that the distance has an impact on power consumption.

5.2.1 Varying Distance between Hub and Relay

Figure 5.8 represents the effective energy in Joules versus the distance between $n_n$ and $n_h$ together with the optimal $\lambda$ values for C-OMA and C-NOMA. The network is composed of $K=5$ nodes by which the relay node’s location varies from $\ell_h^n = 10$ cm to $\ell_h^n = 60$ cm. The other nodes, then, are situated between 70 cm to 100 cm from $n_n$ with an increment of 10 cm between consecutive nodes. Both results suggest that C-NOMA scheme outperforms C-OMA scheme by consuming less energy and
reducing latency in the first phase. From this we deduce that the channel gain of the relay node plays a vital role in power consumption. Additionally, the energy consumption is linked to the phase time allocation $\lambda$. This is because in C-OMA more power is allocated to low channel gain nodes, i.e., first phase nodes. Thus, an increase in the transmission in first phase causes the source nodes to communicate with $n_n$ with higher power. Which justifies the higher energy consumption.

![Figure 5.8: Effect of $\ell_n^h$: a) $E_{eff}$ vs. $\ell_n^h$ and b) Optimal $\lambda$ vs. $\ell_n^h$.](image)

### 5.2.2 Varying Distance between Relay and Source Nodes

Similarly, Figure 5.9 plots the same performance metrics, however, in this case we fix the distance between $n_n$ and $n_h$ to $\ell_n^h = 10$ cm and we change the locations of the source nodes in reference to $n_n$. We achieve this by changing the first source node’s location which has a domino effect on the position of the other source nodes. As we separate consecutive nodes by $\Delta = 10$ cm. Likewise, this figure confirms the effect of source nodes’ distances on the energy consumption. Moreover, C-NOMA exceeds C-OMA and the difference in their performances can reach up to $\Delta E_{effect} = 2.9$ nJ for $\ell_n^{n_1} = 100$. 
5.2.3 Energy Versus number of Nodes

Finally, to assess the performance of the proposed cooperative schemes with regards to number of IoB nodes, Figure 5.10a illustrates the effective energy in Joules versus the number of nodes. The figures shows that the proposed C-NOMA significantly surpasses C-OMA as the number of users increases. This can be explained by the fact that, as the number of source nodes increases, C-OMA assigns higher powers to the first phase over a longer transmission time as seen in Figure 5.10b. Which results in a noticeable gain in favor of C-NOMA.

5.2.4 Regular NOMA versus C-NOMA

Lastly, we compare the performance of regular and cooperative NOMA schemes from an energy consumption perspective with respect to QoS requirements. The network is composed of $K=5$ nodes by which the relay node is located at $\ell^n_h = 10$ cm and $k^{th}$ source node is at $\ell^n_k = k \times 10$ cm. It is obvious from Figure 5.11 that the regular NOMA showcases a reduced energy consumption in contrast to C-NOMA. This due to the fact that in both schemes, the highest transmission power weight is allocated
to the node with the highest gain. In cooperation, the highest gain, i.e., relay node $n_n$ broadcasts the signals of all nodes in the phase of cooperation. Hence, the power coefficients are obtained from the highest gain.

Figure 5.10: Effect of $K$: a) $E_{eff}$ vs. $K$ and b) Optimal $\lambda$ vs.$K$.

Figure 5.11: Regular NOMA vs. C-NOMA
Chapter 6

Conclusion and Future Work

IoB is the awaiting new technology set to revolutionize multiple sectors and industries to ameliorate the quality of life. With the aid of advancements in electronics and wireless technologies, miniaturized wearable sensors can be deployed invasively or non-invasively to monitor, control, and exchange data with authorized off-body entities. However, the scalability of IoB imposes a set of challenges, including meeting QoS demands, energy efficiency, coexistence and interference problems, security, and privacy concerns. Further, adopting pre-existing RF solutions in IoB is nonoptimal due to their power-hungry, bulky, and susceptible to interference modules as well as inherent security vulnerabilities in PHY layer. This prompted the development of energy efficient BCC systems, which confine the signal to the body resulting in an innate PHY security.

In this thesis, we initiate our discussion with a review of the pioneering work that facilitated BCC. We identify the main coupling methods, present the different approaches for optimizing the physical layer, and discuss efforts to enable communication in IoB through BCC. From this, we deduce that networking prospects in BCC require further investigation. Particularly, power constraints and limited battery capacity remain a major hurdle against reliable and long-standing networks. Therefore, we take the initiative and propose ultralow-power OMA and NOMA body channel access schemes to permit multipoint communication. The merits of the proposed schemes are addressed through simulations presented in Chapter 5. It was found the NOMA schemes (both regular and cooperative) are more befitting for IoB systems in
terms of capacity and power consumption.

There are plenty of improvements that could be done in order to achieve results reflecting real-life scenarios. Firstly, we plan to investigate the non-uniform deployment of sensor nodes and the impact on the network lifetime. Because good sensor deployment should maintain high connectivity between the nodes with low energy consumption. We also plan to extend our results for power control levels and maximum feasible number of users to unequal SINR thresholds and imperfect cancellations, respectively. Additionally, we wish to evaluate the robustness of our IoB network in the presence of ambient RF radiations. Intrinsically, BCC confines the signal to a close vicinity to the body, i.e., < 5 cm, which improves interference mitigation. Nevertheless, there’re other wearable devices such as an earphone, smartwatch, and smartphone which exist in close proximity to the body and use RF to communicate. Moreover, we are interested in addressing noises that may arise due to the motion of the body or noises from biological signals. Lastly, we would like to leverage the deployment of IoB networks by integrating networking and circuitry. Hence, addressing implementation holistically.
REFERENCES


APPENDICES

A Papers Submitted and Under Preparation

- Abeer AlAmoudi, Abdulkadir Celik, and Ahmed Eltawil, "Energy Efficient Capacitive Body Channel Access Schemes for Internet of Bodies", Submitted to 2021 IEEE Global Communications Conference (GLOBECOM). This work is discussed in Chapter 3. Available: https://repository.kaust.edu.sa/handle/10754/669768