Outlooks on Radio Transmitter Energy Efficiency and Ultra-Low Power Radio Transmitters

Mika Pulkkinen
Outlooks on Radio Transmitter Energy Efficiency and Ultra-Low Power Radio Transmitters

Mika Pulkkinen

A doctoral thesis completed for the degree of Doctor of Science (Technology) to be defended, with the permission of the Aalto University School of Electrical Engineering, at a public examination held at the lecture hall TU1 of the school on 10 April 2024 at 12:00.

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Abstract

The number of wireless electronic devices in the world continues to grow at a rapid pace. The wireless devices are powered by limited energy sources and require energy-efficient electronic circuitry to maximize the operating time. Radios can consume a significant share of the power budget of a wireless device. Researchers and designers have, therefore, innovated numerous ultra-low power (ULP) radio transmitters and receivers. Such ULP transmitters have conventionally used mostly on-off keying (OOK), binary phase-shift keying (BPSK) or binary frequency-shift keying (BFSK). These modulation schemes enable low power consumption – even below 100 μW – as they only require low-complexity transmitter circuitry and a relatively low signal-to-noise ratio (SNR) for a given error probability.

This thesis investigates the option of using M-ary pulse-position modulation (PPM) and differential PPM (DPPM) in ULP narrowband transmitters in lieu of the conventionally used modulation schemes. The goal is to evaluate whether or not they could achieve better energy efficiency. The benefits and disadvantages of PPM and DPPM are reviewed and their energy efficiency is compared with OOK, BPSK and BFSK in a new way. In ULP transmitters, the power amplifier and carrier synthesizer generally consume the most power. The new comparison method considers how the choice of modulation impacts the combined energy consumed by these blocks per bit. The results suggest that use of OOK, BPSK and BFSK can consume tens to hundreds of percents more energy per bit compared to PPM and DPPM. The M-ary PPM schemes are predicted to be particularly energy-efficient in low-output-power transmitters.

To evaluate the energy efficiencies of transmitter implementations, a new figure of merit (FOM) is derived. In prior ULP transmitter publications, the energy efficiency FOMs have neglected the effect of noise bandwidth and the choice of modulation. However, it is practically the output power and noise bandwidth that together determine the SNR of the generated signal. Moreover, this SNR and the SNR required by the modulation scheme significantly impact the achievable uplink range. By accounting for these metrics in addition to the energy consumption per bit (EPB), an FOM is obtained that provides a more comprehensive view of transmitter energy efficiency than the comparison metrics that have been used before.

Two ULP DPPM radio transmitter implementations are presented with measured results. The first one achieves one of the lowest EBPs while still enabling an uplink range of 30 meters. The second one consumes more energy per bit but enables an uplink range of up to 1 km. An FOM comparison with prior transmitters, including recent sub-mW Bluetooth Low Energy transmitters, suggests that the latter transmitter is state of the art in terms of energy efficiency. In addition to the transmitters, an experimental ULP capacitive gesture sensor interface is presented with measured results. It enables hand-sweep and push-gesture detection over a short range.

Keywords RF, radio, transmitter, energy efficiency, low-power, gesture sensor
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Mika Pulkkinen

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Preface

The main topic in this doctoral thesis is the energy efficiency of radio transmitters. The field of this work is integrated circuit (IC) design and, thus, transmitter circuits and also gesture sensor interface circuits\(^1\) are discussed. Nonetheless, for a thesis in the field of IC design, this work also includes somewhat much modulation-related content. The energy efficiency of a transmitter is significantly impacted by the utilized modulation scheme and its performance. This has often not been fully acknowledged in transmitter IC publications and, particularly, when transmitter energy efficiencies have been analyzed and compared numerically. This has been one of the main motivators for discussing digital modulation techniques extensively in this thesis.

I have tried to present the energy-efficiency-related modulation discussion in sufficient detail. The goal was that this content could be understood without a frequent need to search for pieces of information elsewhere. Nevertheless, as the explanations still are not all-encompassing, I would like to mention as a reading tip that the book reference [7] is, at least at the moment of writing this, available online for IEEE members at https://ieeexplore.ieee.org/book/5271273.\(^2\) The book could quickly provide further modulation-related information. Some modulation-related concepts may be a bit difficult to grasp only by reading. Thus, as another tip, I would like to mention that performing waveform-level receiver simulations can help one to achieve a more profound understanding of, for example, 1) how digital communication using the different modulation schemes works, 2) how and why transmitter energy efficiency is impacted by the utilized modulation scheme, and 3) some signal-to-noise ratio definitions used in communications literature and also in this work. Section 3.3 contains some simulation-related discussion which could be helpful related to that.

This thesis is based on research work which was conducted mostly at the De-\(^1\)Demo available at http://tinyurl.com/gesture-sensor/ and in Youtube with the video title “Ultra-Low Power 2-Axis Gesture Sensor Interface in 180nm CMOS”, posted by “Aalto Electronic Circuit Design”. Filmed using the sensor of publication [V].
\(^2\)One may need to sign in to IEEE Xplore using their personal account to access the book. The book’s availability, when using institutional access, depends on the institution.
partment of Electronics and Nanoengineering, School of Electrical Engineering, Aalto University. I was employed there as a doctoral candidate between 2014 and 2020 and, during this time, I completed most parts of the work reported in this thesis. I continued a part of the modulation-related studies and writing tasks, however, until 2023 besides my job in the industry. The thesis-related work at the university was funded by 1) Aalto ELEC Doctoral School, 2) The Naked Approach and Towards Digital Paradise projects granted by Business Finland, and 3) EffiNano project granted by the Aalto University School of Electrical Engineering. Additional funding was received from Nokia Foundation. I am grateful to the aforementioned funders for supporting this work.

I would like to thank the preliminary examiners, Prof. Patrick P. Mercier and Prof. Zhihua Wang, for reviewing this thesis and for their encouraging feedback and suggestions for improving the content. Hopefully the consequent modifications have improved the quality of the manuscript. I would also like to thank Dr HDR Andreia Cathelin for accepting the role as the opponent of this work. It is a major honor to have such a distinguished group of professionals as the reviewers. Thank you, all, for taking the time to read this manuscript.

This thesis work was supervised by Prof. Kari Halonen. I would like to thank Prof. Halonen for his guidance, leadership and collaboration with the publications. Especially, I would like to thank him for welcoming me to the Electronic Circuit Design (ECD) unit in 2008, for the versatile work tasks he assigned to me during the years, and for the related work experience I gained. The IC design field is extremely multifaceted and it has been a privilege to get to experience so many sides of it. I could not have wished for a more educational start for my career. Thank you, Prof. Halonen, for the opportunity to learn so much. Moreover, thank you for being an excellent superior and, particularly, for being compassionate when the times called for that.

In the thesis-related research projects, I worked closely with Dr. Tuomas Haapala, Jarno Salomaa and Mohammad Mehdi Moayer. We collaborated related to the IC designs and also partly related to the doctoral studies. Jarno, Tuomas and Mehdi, thank you for the inspiring conversations and brainstorms, the extracurricular activities and the unforgettable tape-outs. Designing circuits and conducting research with people like you – that is my definition of a dream job. In the ordinary days, I will keep missing, among other things, your sense of humor and our moments by the whiteboard (but luckily I get to enjoy at least the humor every now and then). Thanks, Jarno and Tuomas, for sharing some of your analog and RF wisdom with me during my transition from designing digital circuits more towards designing analog and RF ones. Also, Tuomas, thanks for your tips related to this thesis process. Doru Irimescu, thanks for assisting us with the design of the PCB for the system-on-chip measurements.

For their teachings and assistance, I would like to say thank you to the following long-term staff members and cornerstones of the research group: Prof. Jussi Rynänen, Dr. Marko Kosunen and Dr. Kari Stadius. Artturi Kaila, thanks for all the help related to IT, and thanks for lending me the old Casio
keyboard – that favor still positively impacts my life. Dr. Mikko Englund, thank you for our discussions at the ECD unit and thanks for establishing a connection between me and CoreHW (in 2020, it became my next and current employer).

I was with the ECD unit altogether from 2008 to 2020. The premises of the research group always felt like a second home with a pleasant and extremely educational atmosphere. For that, I am grateful to numerous past and current members of the group including the ones mentioned in the previous paragraphs. Looking back at those years, I can recall many many many friendly, bright, inspiring and instructive coworkers. There are perhaps a few tens of persons too many for thanking everyone here individually. Thus, Prof. Halonen and all the named and unnamed former and present members of the ECD unit who were around for a shorter or longer time between 2008 and 2020, please, accept my humble group thank-you: thanks for creating the outstanding workplace and its atmosphere, thanks for everything I learned with you and from you, and thanks for the warm and fun memories related to things such as bowling, coffee breaks at Otakaari 5, conferences, floorball, Friday infos, IC design, ice swimming, karonkkas, lunches, measurement lab, recreation days, Red Alert, STIGA hockey, table football, tape-outs, teamwork and work trips. Generally, big thanks for the unforgettable experience which it was to study the art and science of IC design in such great company. I would additionally like to thank all of the former and current secretaries, HR personnel and other staff on ECD-, department- and school-level who have helped me with the practical tasks related to the doctoral study process and the work at the university.

Finally, I would like to express my gratitude to my closest ones. I would like to thank my loving mother and father, Kirsti† 7 July 2023 and Kalervo, for their encouragement and support throughout this educational path since the first day of school. Thank you, my sister, Katja, for being near and for helping to overcome the downsides of life. The last few years have been laborious and a bit too busy. For delighting this life during these busy years, I would like to thank my family, my beloved one, Isabella, our tiny furry friend, Milo, and all my dear friends and bandmates. Although this thesis work has been extremely interesting, I also look forward to finishing this process and having more time to spend with you.

Espoo, March 19, 2024,

Mika Pulkkinen
Preface

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List of Publications

This thesis consists of an overview and of the following publications which are referred to in the text by their Roman numerals.


Author’s Contribution

Publication I: “Outlooks on transmitter energy efficiency and FOM and a –189.7-dBJ/bit ULP DPPM transmitter”

The author developed the method for comparing the energy efficiencies of modulation schemes which accounts for the power consumed by both a power amplifier and a carrier synthesizer. The author derived the new figure of merit for radio transmitter energy efficiency. The author designed 1) the narrowband transmitter consisting of the power amplifier, local oscillator, modulator circuitry and ring oscillator, 2) the gesture sensor interface and the related digital signal processing block, and 3) the digital control circuitry for the chip including the serial peripheral interface. The author performed all the reported measurements and wrote the manuscript.

Publication II: “Low-power wireless transceiver with 67-nW differential pulse-position modulation transmitter”

The author designed 1) the narrowband transmitter including the power oscillator, DPPM modulator and on-PCB transmit antenna, 2) the temperature-compensated ring oscillator, comparator and digital back-end for the receiver, and 3) the digital control circuitry for the chip including the serial peripheral interface. The author also designed the measurement PCB, performed all the measurements and wrote the manuscript excluding Section V-A which discusses the receiver front-end.
Publication III: “Low-power single-stage narrowband transmitter front-end for 433-MHz band”

The author designed the power oscillator and the on-PCB transmit antenna. The author also designed the measurement PCB, performed all the measurements, wrote the manuscript and presented the paper at the conference.

Publication IV: “45.2% energy efficiency improvement of UWB IR Tx by use of differential PPM in 180nm CMOS”

The author wrote the VHDL codes of the digital modulator blocks, performed the related synthesis and place-and-route, designed the ring oscillator, performed the measurements excluding the pulse generator characterization, wrote the manuscript and presented the paper at the conference.

Publication V: “462-nW 2-axis gesture sensor interface based on capacitively controlled ring oscillators”

The author designed the capacitance-controlled ring oscillator, reference ring oscillator and digital blocks of the gesture sensor interface, performed the measurements, wrote the manuscript and presented the paper at the conference.
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# List of Abbreviations

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<th>Abbreviation</th>
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<td>ADC</td>
<td>analog-to-digital converter</td>
</tr>
<tr>
<td>ASK</td>
<td>amplitude-shift keying</td>
</tr>
<tr>
<td>AWG</td>
<td>additive white Gaussian</td>
</tr>
<tr>
<td>AWGN</td>
<td>additive white Gaussian noise</td>
</tr>
<tr>
<td>BCG</td>
<td>bias current generator</td>
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<tr>
<td>BER</td>
<td>bit error ratio</td>
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<tr>
<td>BFSK</td>
<td>binary frequency-shift keying</td>
</tr>
<tr>
<td>BLE</td>
<td>Bluetooth Low Energy</td>
</tr>
<tr>
<td>BPF</td>
<td>bandpass filter</td>
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<tr>
<td>BPSK</td>
<td>binary phase-shift keying</td>
</tr>
<tr>
<td>CCRO</td>
<td>capacitance-controlled ring oscillator</td>
</tr>
<tr>
<td>CDF</td>
<td>cumulative distribution function</td>
</tr>
<tr>
<td>CIC</td>
<td>cascaded integrator-comb</td>
</tr>
<tr>
<td>CMOS</td>
<td>complementary metal-oxide-semiconductor</td>
</tr>
<tr>
<td>CNR</td>
<td>carrier-to-noise ratio</td>
</tr>
<tr>
<td>DAC</td>
<td>digital-to-analog converter</td>
</tr>
<tr>
<td>dBm</td>
<td>decibels relative to one milliwatt</td>
</tr>
<tr>
<td>DCO</td>
<td>digitally controlled oscillator</td>
</tr>
<tr>
<td>DE</td>
<td>drain efficiency</td>
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<tr>
<td>DPPM</td>
<td>differential pulse-position modulation</td>
</tr>
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<td>DSP</td>
<td>digital signal processing</td>
</tr>
<tr>
<td>EC</td>
<td>edge-combiner</td>
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<tr>
<td>ECPA</td>
<td>edge-combining power amplifier</td>
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<tr>
<td>ED</td>
<td>envelope detector</td>
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<td>EPB</td>
<td>energy consumption per bit</td>
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<td>ETSI</td>
<td>European Telecommunications Standards Institute</td>
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<tr>
<td>FBAR</td>
<td>film bulk acoustic resonator</td>
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<tr>
<td>FOM</td>
<td>figure of merit</td>
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<tr>
<td>FPGA</td>
<td>field-programmable gate array</td>
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<tr>
<td>FSK</td>
<td>frequency-shift keying</td>
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<tr>
<td>GE</td>
<td>global efficiency</td>
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<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>GFSK</td>
<td>Gaussian frequency-shift keying</td>
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<td>GMSK</td>
<td>Gaussian minimum-shift keying</td>
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<tr>
<td>GS</td>
<td>guard slot</td>
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<tr>
<td>HP</td>
<td>high-power</td>
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<tr>
<td>IC</td>
<td>integrated circuit</td>
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<tr>
<td>IF</td>
<td>intermediate frequency</td>
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<tr>
<td>ILRO</td>
<td>injection-locked ring oscillator</td>
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<tr>
<td>IoT</td>
<td>Internet of Things</td>
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<tr>
<td>ISM</td>
<td>industrial, scientific and medical</td>
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<td>LF-LO</td>
<td>low-frequency local oscillator</td>
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<td>LO</td>
<td>local oscillator</td>
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<tr>
<td>LOS</td>
<td>line-of-sight</td>
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<tr>
<td>LP</td>
<td>low-power</td>
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<tr>
<td>LSB</td>
<td>least significant bit</td>
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<tr>
<td>MIM</td>
<td>metal-insulator-metal</td>
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<td>MOM</td>
<td>metal-oxide-metal</td>
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<td>MSF</td>
<td>moving sum filter</td>
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<tr>
<td>MSK</td>
<td>minimum-shift keying</td>
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<td>MSPS</td>
<td>megasamples per second</td>
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<td>OBW</td>
<td>occupied bandwidth</td>
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<tr>
<td>OOK</td>
<td>on-off keying</td>
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<tr>
<td>PA</td>
<td>power amplifier</td>
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<tr>
<td>PCB</td>
<td>printed circuit board</td>
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<tr>
<td>PDF</td>
<td>probability density function</td>
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<tr>
<td>PER</td>
<td>packet error ratio</td>
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<tr>
<td>PG</td>
<td>pulse generator</td>
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<tr>
<td>PL-SDD</td>
<td>packet-level soft-decision decoding</td>
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<tr>
<td>PLL</td>
<td>phase-locked loop</td>
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<td>PO</td>
<td>power oscillator</td>
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<td>PPM</td>
<td>pulse-position modulation</td>
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<tr>
<td>PS</td>
<td>proximity sensor</td>
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<tr>
<td>PSD</td>
<td>power spectral density</td>
</tr>
<tr>
<td>PSK</td>
<td>phase-shift keying</td>
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<tr>
<td>PV</td>
<td>photovoltaic</td>
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<tr>
<td>PWM</td>
<td>pulse-width modulation</td>
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<tr>
<td>QAM</td>
<td>quadrature amplitude modulation</td>
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<tr>
<td>RBW</td>
<td>resolution bandwidth</td>
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<tr>
<td>RF</td>
<td>radio frequency</td>
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<td>RMS</td>
<td>root mean square</td>
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<td>RO</td>
<td>ring oscillator</td>
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<tr>
<td>RX</td>
<td>receiver</td>
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<tr>
<td>SAW</td>
<td>surface acoustic wave</td>
</tr>
<tr>
<td>SER</td>
<td>symbol error ratio</td>
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<tr>
<td>SNR</td>
<td>signal-to-noise ratio</td>
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<table>
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<th>Description</th>
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<tr>
<td>SoC</td>
<td>system-on-chip</td>
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<tr>
<td>SPI</td>
<td>serial peripheral interface</td>
</tr>
<tr>
<td>SPS</td>
<td>samples per second</td>
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<tr>
<td>TEG</td>
<td>thermoelectric generator</td>
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<tr>
<td>TRX</td>
<td>transceiver</td>
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<tr>
<td>TX</td>
<td>transmitter</td>
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<tr>
<td>ULP</td>
<td>ultra-low power</td>
</tr>
<tr>
<td>UWB</td>
<td>ultra-wide band</td>
</tr>
<tr>
<td>VBW</td>
<td>video bandwidth</td>
</tr>
<tr>
<td>VCO</td>
<td>voltage-controlled oscillator</td>
</tr>
<tr>
<td>VHDL</td>
<td>very-high speed integrated circuit hardware description language</td>
</tr>
<tr>
<td>WLAN</td>
<td>wireless local area network</td>
</tr>
<tr>
<td>WSN</td>
<td>wireless sensor network</td>
</tr>
<tr>
<td>XCP</td>
<td>cross-coupled pair</td>
</tr>
<tr>
<td>XO</td>
<td>crystal oscillator</td>
</tr>
</tbody>
</table>
List of Symbols

\( a \) dilation factor
\( A \) signal envelope
\( A_c \) amplitude of a carrier wave
\( B \) number of bits encoded per symbol
\( BW_N \) noise bandwidth
\( BW_{null} \) null-to-null bandwidth
\( C \) capacitance
\( C_S \) capacitance of a sensor element
\( d \) distance
\( E_b \) signal energy per bit
\( E_b/N_0 \) signal-to-noise ratio per bit, also denoted as \( \gamma_b \)
\( E_s \) signal energy per symbol
\( EPB_{rel,mod} \) EPB with modulation scheme \( mod \) relative to a reference modulation scheme
\( f \) frequency
\( f_{BB} \) baseband clock frequency
\( f_c \) carrier frequency
\( G_{P,mod} \) output power scaling factor with modulation scheme \( mod \)
\( I_B \) bias current
\( I_D \) drain current
\( k \) Boltzmann constant
\( L_{p,max} \) maximum number of slots in a DPPM packet
\( L_{s,max} \) maximum number of slots in a DPPM symbol
\( LS \) link strength
\( N_0 \) noise power spectral density
\( N_O \) number of “on” slots in a DPPM packet
\( N_b \) number of bits transmitted in a packet
\( N_s \) number of symbols in a packet
\( N_Z \) maximum number of “off” slots in a DPPM packet
\( P_{LO} \) power consumption of local oscillator
\( P_n \) noise power
List of Symbols

- $P_{out}$: output power
- $P_{PA}$: power consumption of power amplifier
- $P_{rad, out}$: radiated output power
- $P_{TX}$: total power consumption of a transmitter
- $R_b$: data rate
- $R_{DC}$: duty cycle
- $\gamma_{t, mod}$: active ratio of modulation scheme $mod$
- $T$: noise temperature
- $T_{BB}$: baseband clock period
- $T_s$: symbol time
- $W$: integer value of a data word or a symbol
- $\alpha$: ratio of reference transmitter's PA and LO power consumptions
- $\gamma$: signal-to-noise ratio
- $\gamma_{avg, req}$: average-signal-power-to-noise-power ratio required by a modulation scheme
- $\gamma_b$: signal-to-noise ratio per bit, also denoted as $E_b/N_0$
- $\gamma_{b, req}$: signal-to-noise ratio required per bit by a modulation scheme
- $\gamma_{max}$: maximum SNR achievable with TX output signal
- $\gamma_{req}$: signal-to-noise ratio required by a modulation scheme
- $\eta$: power efficiency
- $\eta_{PA}$: efficiency of power amplifier
- $\sigma$: standard deviation of additive white noise
- $\phi$: carrier phase
1. Introduction

1.1 Background

Radio technology plays a key role in our modern societies. It is one of the technologies that enable wireless data transfer and, for instance, access to the data on the Internet from nearly anywhere – all you need is a pocket-size electronic device equipped with a radio transmitter and a radio receiver. The amount of wireless devices and their applications is growing and new radio-equipped products are released at a steady pace. The current and future applications require radios with different specifications. Some applications require high data rates, others require ultra-low power consumption, and research is ongoing to push these qualities to the extreme. Hence, despite the fact that radio technology is old, transmitters, receivers and communication systems are researched extensively still today, 136 years [1] after the existence of radio waves was first confirmed experimentally.

In a sense, it could be said that we are becoming surrounded by radios. Typical everyday objects that utilize wireless connection are mobile phones, tablets and laptops. Radios are additionally used, for example, in car keys for remote control of the locks or the trunk lid and to detect the presence of the car owner. During recent years, a wireless connection has also been included in wearables and body-worn items such as fitness tracker bracelets, wireless headphones and smart watches, rings, glasses and clothing. Other types of devices equipped with radios are portable speakers, televisions, refrigerators, home lighting systems, home voice controllers and wireless security cameras. The number of different applications where radios are utilized is already large. Nevertheless, this may be just the beginning of an outbreak of radio ubiquitousness, as even more applications are envisioned such as smart contact lenses, wireless sensor networks (WSNs) and various Internet of Things (IoT) devices.

The wirelessness of various electronic everyday objects makes their use more comfortable. On the other hand, it also brings about a familiar problem – the energy source often seems to drain all too rapidly. This is generally caused by
having a device that consumes too much power and energy with respect to the
capacity of the energy source which is typically a battery, a rechargeable battery
or an energy harvester. As a result, the battery needs replacement or recharging
frequently or, if an energy harvester is utilized, the device may need to spend
time in sleep mode to gather sufficient energy for the next operation. These
kinds of power outages are generally undesirable and it is preferable to reduce
their rate of occurrence. This can be achieved by minimizing the power and
energy consumption of the device. In a wireless device, and particularly in a
wireless sensor node, a significant share of the power can be consumed by radios.
Therefore, an interest has grown in the research of ultra-low power radios.

This thesis work was conducted as a part of research projects whose common
objective was to implement circuits for an energy-autonomous (i.e. batteryless)
wireless sensor node. Wireless sensor nodes generally contain several subsys-
tems such as a sensor interface, a transmitter and a receiver. To enable operating
with a limited amount of harvested energy, the power consumption of all the
subsystems must be minimized. For example, a commercial 6.5 cm$^2$ photovoltaic
array may produce only 25 microwatts of power under office illumination [2].
To enable powering using such an array without outages, the average power
consumption of each subsystem should be in the order of a few microwatts. These
kinds of strict power requirements have been the motivator for implementing
the ultra-low power transmitters and gesture sensor interface presented in this
work.

1.2 Energy-autonomous wireless sensor node

Fig. 1.1 shows the wireless sensor system-on-chip (SoC) targeted in this work.
It contains several functional blocks along with circuits for energy harvesting
and power management. The functional blocks contain a capacitive 2-axis
gesture sensor, a universal multi-sensor interface, two radio transmitters, a
radio receiver, and circuits for driving an electrochromic icon display.

The system is powered using a photovoltaic (PV) cell or a thermoelectric
generator (TEG). The power management circuitry can store energy in a su-
percapacitor. With this stored energy, the functional blocks can be powered for
limited periods of time when no energy is available from the environment. The
energy harvesting and voltage regulation block generates supply voltages of 0.9,
1.2, 1.8 and 3.3 V for the functional blocks. Several voltages are used for energy
efficiency – some blocks require a higher supply voltage that would result in
excessive power consumption by such blocks that can cope with lower supply.
For example, the ultra-wide band (UWB) impulse radio transmitter requires a
1.8-V supply to be able to produce fast transients which enables operation in
the UWB band from 6.0–8.5 GHz. The intended electrochromic display device,
on the other hand, requires a supply voltage in the range of 3.3 V. The sensor
interfaces and digital blocks do not require such high supply voltages and utilize
0.9 and 1.2 V for lower power consumption.

The capacitive 2-axis gesture sensor interface can be used as a human interface device. It consists of two capacitive proximity sensor (PS) interfaces that provide digital outputs. The use of two PS interfaces connected to two sensor elements enables 2-axis operation. The two interfaces output 24-bit digital data words, each at a sample rate of approximately \( \frac{1}{50} \) samples per second (SPS). The digital outputs are provided to a low-power gesture detection block. Based on this data, the detection algorithm can recognize a push gesture and hand sweeps from left to right and right to left. The gesture sensor interface of this SoC was originally presented in [I] and is discussed in more detail in Chapter 6.

The universal multi-sensor interface can be used with capacitive and resistive sensor elements. For example, temperature, humidity and pressure can be sensed using a proper type of sensor element. The sensor interface converts the capacitance or resistance to a voltage. A 1-bit ΔΣ analog-to-digital converter (ADC) is used for converting this voltage, or the output of a voltage-output sensor, to digital 1-bit data. A digital signal processing (DSP) block decimates the 1-bit data and provides multibit sensor data at a lower sampling rate. The ADC produces output data with an accuracy of 14 bits and the maximum sample rate of the digital data is 10 kSPS. However, it can also be operated in incremental mode to perform sensor measurements with a reduced sample rate, e.g. one

\[ \text{The sampling rate is not accurate because a ring oscillator is used as a reference clock in the frequency-to-digital conversion.} \]
measurement per second or even less often. The universal multi-sensor interface is discussed in reference [3].

Using the radio circuits, this wireless sensor node communicates with a base station device. A base station device here refers to any device that has suitable receiver circuitry, sufficient energy for receiving and processing the sensor data and, perhaps, the means for passing it forward e.g. to cloud. For uplink, two transmitters have been implemented: a 434-MHz narrowband transmitter and the UWB transmitter. The two transmitters offer different uplink ranges. The peak output power of the narrowband transmitter is –2 dBm which enables a high uplink range even with a moderately wide signal bandwidth\(^2\) in the range of a few MHz. However, the energy consumed per transmitted bit is moderate. The UWB transmitter, on the other hand, consumes less energy per bit but offers a reduced uplink range. The range is lower because the transmit power of UWB transmitters is more strictly regulated and generally only low output power spectral density (PSD) is allowed. Moreover, path loss is higher with the higher carrier frequency of the UWB transmitter. Thus, the uplink range is much lower with the UWB transmitter. Either one of the transmitters can be used depending on the required uplink range. Section 5.2 discusses the narrowband transmitter, originally presented in [I]. An optional narrowband transmitter design [II] is discussed in Section 5.1. Additionally, Section 5.4 presents an early prototype of the UWB transmitter which has been published in [IV] and [2].

The gesture sensor and the universal multi-sensor interface define the data rate requirements for the transmitters. The highest data rate is required by the ΣΔ ADC output data of the multi-sensor interface: streaming all data samples requires\(^3\) a rate of 140 kbps. A lower data rate of 2.4 kbps suffices for transmitting the gesture sensor data using full word length.\(^4\) Moreover, the accuracy of the sensor data is lower than 24 bits and the data can also be truncated, for example, to 10 bits. In that case a data rate of 1 kbps is sufficient. In some use scenarios, the actual transmit data rate can also be extremely low. The transmitters could be utilized to transmit data only when some specific event occurs in the sensor data. The multi-sensor interface could be used, for instance, to detect a rise in temperature, or the gesture sensor could be used to detect a sweep of a user's hand. Such events can be rare and, consequently, the data rate can be significantly lower than one bit per second on average. Thus, the transmitters should enable a data rate of 140 kpbs but also consume only minor quiescent power for efficient operation in event-based use scenarios.

The downlink utilizes a 434-MHz narrowband receiver that receives Manchester-encoded data. Manchester encoding is utilized as the encoded data can be received using a noncoherent envelope-detecting receiver front-end without sym-

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\(^2\)The achievable uplink range is determined not only by the output power, but also the power of the noise in the noise bandwidth and the utilized modulation scheme. These topics will be discussed throughout Chapters 2 to 5.

\(^3\)14-bit samples produced at a sample rate of 10 kSPS.

\(^4\)Two 24-bit data streams produced at a sample rate of 50 SPS.
bol synchronization. This type of receiver can achieve a low power consumption and a decent sensitivity. The receiver details are presented in [II].

The SoC additionally comprises circuits for driving an electrochromic icon display. The display can be used as a visual indicator. Its state can be changed, for example, based on the sensor data to indicate that temperature, pressure or humidity has crossed a certain threshold. Alternatively, the display could be used to provide visual feedback to a user when a gesture command has been detected.

1.3 Objectives of the work

This thesis work has contributed to the design of the transmitter circuits and the gesture sensor interface of the SoC presented in the previous section. The objectives of this work were as follows.

1.3.1 Ultra-low power transmitters

The objective of the transmitter part of this work was to research and develop energy-efficient ultra-low power radio transmitters. The targeted application was the energy-harvesting sensor node presented in Section 1.2. This set requirements for the implemented transmitters: 1) power consumption must be in the order of a few microwatts, and 2) they must be capable of transmitting data at rates up to 140 kbps. Low-power transmitters have been researched extensively throughout the last two decades. The goal of this work was to find transmitter solutions that could improve the energy efficiency compared to earlier works.

1.3.2 Low-power gesture sensor interface

In the gesture sensor part of this work, the goal was to implement an experimental ultra-low power gesture sensor prototype and characterize its performance. The prototype is based on ring oscillators and consumes microwatt-range power. The goal was to examine at what range it could detect simple hand gestures such as a sweep and a push.

1.4 Contents and organization of the thesis

The main focus of this thesis is on ultra-low power (ULP) narrowband radio transmitters and radio transmitter energy efficiency. Chapter 2 gives an overview of prior sub-mW transmitter publications. The beginning of the chapter discusses popular digital modulation techniques that have been used in published ULP transmitters: on-off keying (OOK), binary phase-shift keying (BPSK) and binary frequency-shift keying (BFSK). The choice of modulation scheme defines what
kind of a signal a transmitter must generate and can have a significant effect on the power consumption and energy efficiency of a transmitter. OOK, BPSK and BFSK have been popular because the modulated RF signals can be generated with low-complexity circuits using low power.

Low power consumption is naturally desirable in wireless devices but the power consumption alone is not a good metric of transmitter performance. A brief review is given of the transmitter energy efficiency metrics that have been utilized in prior publications. The utilized metrics have practically neglected the signal-to-noise ratio (SNR) and the choice of modulation despite the fact that these are critical properties related to uplink performance. A new comprehensive figure of merit (FOM) for transmitter energy efficiency is then derived that accounts for these properties in addition to consumed energy per bit. The new FOM, originally presented in [I], enables a fairer energy efficiency comparison between transmitters that may use different modulation schemes and differ in terms of output power, consumed power and data rate. After the FOM discussion, a brief review is provided of the state-of-the-art ULP transmitter architectures, carrier signal generation methods and power amplifiers (PAs). In the end of Chapter 2, some transmitter publications are discussed in detail and the achieved performance are compared including the energy efficiency FOMs.

Chapter 3 is based on [I] and [III]. It discusses two alternative modulation techniques for use in lieu of the conventionally used OOK, BPSK and BFSK: M-ary pulse-position modulation (PPM) and M-ary differential pulse-position modulation (DPPM). M-ary PPM possesses various favorable properties in terms of energy efficiency. One major benefit is that, due to orthogonality, PPM is energy-efficient in terms of SNR. With an orthogonal signal, the more bits are encoded per symbol, the less SNR is required per bit to achieve a given symbol error ratio (SER) or bit error ratio (BER). This enables a reduced PA energy consumption per bit (EPB). Another benefit is evident when multiple bits are encoded per symbol: the more bits are encoded per symbol, the lower is the time that the carrier is transmitted per bit. This too can reduce the energy consumption per bit. The key feature with PPM is that only a minor output power increment is required to maintain error performance when the number of bits per symbol is increased. Therefore, PPM enables encoding of multiple bits per symbol with relatively low output power for a given error probability. Additionally, PPM transmission can be performed using similar low-complexity low-power transmitter architectures that have been used in prior ULP transmitters. A PPM modulator itself is a relatively low-complexity circuit as well – it can be implemented as a counter-based digital circuit that can consume very low power compared to the local oscillator (LO) and PA. With PPM, increasing the number of bits encoded per symbol requires little added complexity to the TX circuitry – mainly the word length of the digital modulator must be increased. These properties make PPM an advantageous modulation for ULP transmitters.

The second modulation scheme discussed in Chapter 3, DPPM, offers many
of the same benefits as PPM. A DPPM signal resembles a PPM signal and can be generated with similar low-complexity low-power transmitter circuitry and a digital modulator. In a similar manner with PPM, multiple bits can be encoded per transmitted RF pulse for decreased active time of the local oscillator and power amplifier per bit and, consequently, for improved energy efficiency. In the same fashion with PPM, DPPM requires only a minor output power increment for maintaining the error performance when the number of bits per symbol is increased. A DPPM signal differs from a PPM signal, however, as each DPPM symbol ends immediately after the RF pulse. For this, the data rate is greater than with PPM but the signal is not orthogonal and the error performance is degraded. Nonetheless, a soft-decision decoding method [II] can be used that achieves a good error performance that even outperforms BPSK. The decoding method is described in the chapter. Due to these benefits, DPPM is a considerable modulation for ULP transmitters as well.

In Chapter 4, PPM and DPPM are compared with the conventionally used binary modulation schemes in terms of circuit energy consumption. An analysis is provided of the effect of the choice of modulation scheme on the combined energy consumed per bit by a carrier synthesizer and a PA. These two blocks can often consume the most power in ULP transmitters and, thus, the focus is on them. Different modulation schemes require different SNRs to achieve a given error probability. This implies that, for equal error performance, the output power needs to be adjusted depending on the chosen modulation scheme. This affects the power and energy consumption of the PA. The LO and PA energy consumptions per bit are furthermore affected by the number of bits encoded per symbol. With more bits encoded per symbol, the transmitter is active for less time per bit which results in a reduced energy consumption per bit. These effects are considered in the analysis. The results suggest that a transmitter consumes less energy per transmitted bit if M-ary PPM or DPPM is used instead of OOK, BPSK or BFSK. The presented analysis method could also be used for assessing the energy efficiency of other modulation schemes than those which are considered in this thesis. The analysis method and the key results have originally been presented in [I].

Chapter 5 presents two narrowband transmitters which were implemented as a part of this thesis work and presented in [I] and [II]. The prototypes support DPPM, performed using on-chip modulators, and achieve state-of-the-art performance in terms of power consumption and energy efficiency. The prototype described in Section 5.1 achieves one of the lowest EPBs and power consumptions among ULP narrowband transmitters and still achieves a measured line-of-sight uplink range of 30 meters [II]. The prototype of Section 5.2 consumes greater energy per bit but is also expected to achieve a longer line-of-sight uplink range of up to 1 km [I]. A comprehensive energy efficiency comparison is made with other ultra-low power transmitters including sub-mW Bluetooth Low Energy (BLE) transmitters. The comparison suggests that the second prototype is state of the art in terms of energy efficiency. In addition to the narrowband
transmitters, the end of the chapter discusses the EPB reduction achieved in [IV] by use of DPPM in the case of the UWB transmitter.

In Chapter 6, the basics of capacitive proximity sensors are discussed briefly. The prototype 2-axis gesture sensor interface of [I] is presented. The sensor consumes microwatt-range power and enables detection of hand sweep and push gestures at a distance of 12 cm. Finally, Chapter 7 concludes the thesis.

1.5 Main scientific merits

The scientific contributions of this work are presented in detail within Publications [I]–[V]. A summary of the work in these publications is found in Chapters 2–6 of this thesis. The most important scientific contributions of this work can be summarized as follows:

1. The packet error ratio (PER) with DPPM was analyzed [II] considering a scenario where a packet-level soft-decision decoding (PL-SDD) scheme is utilized. PER comparison was performed with on-off keying (OOK) [II]. The results suggest that DPPM enables a better error performance than OOK except at very low values of SNR $\gamma$. Thus, use of DPPM enables lower energy consumption per bit compared to OOK without deteriorating error performance.

2. The benefits of utilizing M-ary modulation schemes in ultra-low power narrowband transmitters were addressed [I], [II]. ULP narrowband transmitters have generally utilized binary modulation schemes. However, the use of M-ary modulation schemes enables a lower active time per bit of high-power transmitter blocks. This can reduce the energy consumed per transmitted bit. In this work, particular focus has been on PPM and DPPM that require a low signal-to-noise ratio per bit to achieve a given error probability. Because of this property and the low active time per bit, they enable low energy consumption per bit by both a power amplifier and a carrier synthesizer.

3. Modulation schemes that have been popular in ULP transmitters were compared in a new way [I]. In ULP transmitters, the output power can be low and power consumption is not necessarily significantly dominated by the PA. For this, the energy efficiency of a modulation scheme is not solely determined by the SNR it requires per bit. Consequently, it is reasonable to consider how the choice of modulation scheme impacts the energy consumed by also other power-consuming blocks in addition to the PA. In ULP transmitters, a significant amount of power is also consumed by carrier synthesis. For this, the new method considers how the choice of modulation scheme affects the combined energy consumed per bit by the LO and the PA when the output power with each modulation scheme is scaled for equal uplink strength. In [I], OOK, BPSK and BFSK were compared with PPM and DPPM. The results
suggest that PPM and DPPM could enable lower energy consumption per transmitted bit than OOK, BPSK and BFSK. The M-ary PPM and DPPM schemes are estimated to be particularly energy-efficient in low-output-power scenarios.

4. A new figure of merit was derived for radio transmitter energy efficiency [I]. In contrast to earlier FOMs that have been used to characterize low-power transmitters, the new FOM accounts for the maximum signal-to-noise ratio achievable with the output signal and the SNR required by the utilized modulation scheme. By accounting for the noise bandwidth through the SNR, the FOM avoids improper data rate dependency. Generally, by accounting for metrics that have been neglected in earlier energy FOMs for transmitters, the new FOM enables a more rational energy efficiency comparison between transmitters that differ in terms of consumed power, output power, data rate and modulation.

5. An ultra-low power 434-MHz narrowband DPPM transmitter [II] was designed in 180 nm CMOS. It consumes energy down to 11.6 pJ/bit and the peak output power is approximately –25 dBm. The lowest power consumption is 8.3 μW when data is transmitted continuously at a data rate of 17.9 kbps using a baseband clock frequency of 100 kHz. The measured power consumption is 67 nW when data is transmitted in packets at a data rate of 4.8 kbps using a higher baseband clock frequency of 5 MHz, i.e. DPPM pulses with a shorter duration. In the measurements, the data transmitted in this mode was received at a 30-meter distance from the transmitter using a non-directional receiving antenna. With narrower signal bandwidth and higher EPB, the line-of-sight uplink range is up to 200 m. The achieved power consumption and energy consumption per bit during continuous-mode data transmission are among the lowest ones achieved by published state-of-the-art narrowband transmitters.

6. Another ultra-low power 434-MHz narrowband DPPM transmitter [I] was designed in 180 nm CMOS. It consumes energy down to 0.52 nJ/bit when data is transmitted continuously. When DPPM data is transmitted in packet-mode 48 bits/packet at a data rate of 2.27 kbps, the transmitter consumes 1.56 μW of power and 0.69 nJ/bit. The peak output power is –2 dBm and the estimated line-of-sight uplink range is up to 1 km. The power efficiency and energy efficiency FOM are among the best ones achieved by published state of the art.

7. A DPPM modulator [IV] was designed for an ultra-wide band transmitter. It was shown that, with the utilized circuit implementations, using DPPM instead of OOK decreases the energy consumption of the transmitter per bit by up to 56%.

8. Two versions [I], [V] of a low-power capacitive 2-axis gesture sensor interface
were designed and implemented and the performance was characterized. Using 462 nW of power, the first version [V] enables detection of hand sweep gestures at a distance of 6 cm. The second version [I] consumes 3.2 µW of power and enables gesture detection at a distance of 12 cm.
2. Binary modulation schemes, literature review and new FOM for transmitter energy efficiency

This chapter discusses typical modulation schemes that are utilized in published state-of-the-art ULP transmitters and, furthermore, the circuits required for generating the modulated signals. Some essential basic concepts of modulation are revised briefly. Additionally, some aspects regarding transmitter energy efficiency are discussed. A new concept named the active ratio of a modulation scheme is introduced at the end of Section 2.1.1, and determined for all the modulation schemes whose energy efficiencies are compared later in this work. More information about the topics and other optional modulation schemes can be found in [I] and [II] in addition to references [1], [4], [5], [6] and [7]. At the end of the chapter, ULP narrowband transmitter architectures, their key building blocks and some state-of-the-art transmitters are reviewed.

Modulation is a signal processing operation where some parameter of a carrier wave is changed in accordance with an information-bearing signal [1]. In radio communications, the carrier is a sinusoidal wave. The equation of a sinusoidal carrier wave is

\[ c(t) = A_c \cdot \cos(2 \cdot \pi \cdot f_c \cdot t + \phi), \]

where the parameters \( A_c, f_c \) and \( \phi \) are the amplitude, frequency and phase of the wave. In digital radio communications, modulation can be performed by toggling any of these parameters based on the transmit data. Correspondingly, the basic types of digital modulation are amplitude-shift keying (ASK), frequency-shift keying (FSK), and phase-shift keying (PSK) [5]. Alternatively, modulation can be performed by multiplying the RF carrier wave with a periodic sequence of pulses whose width or position is modified according to the transmit data [1]. These types of modulation schemes are, for instance, pulse width modulation (PWM), pulse-position modulation (PPM) and differential pulse-position modulation (DPPM) [6], [8], [9]. Moreover, different modulation methods can be combined. For example, with quadrature amplitude modulation (QAM), data is encoded simultaneously to the amplitude and phase of the transmitted signal.

Within published ULP narrowband transmitters, binary ASK, PSK and FSK have been popular, binary meaning that one bit is encoded per symbol. Within those ULP transmitters that consume less than 1 mW of power during trans-
mission, these modulation schemes clearly dominate\(^1\). With these modulations, the modulator is clocked with the baseband clock frequency \(f_{BB}\) and one bit is transmitted per symbol in one baseband clock cycle \(T_{BB} = 1/f_{BB}\). For each duration of \(T_{BB}\), the amplitude, phase or frequency is toggled to either of two possible values that represent the transmit bit whose value is 0 or 1.

### 2.1 Popular modulation schemes in published ULP transmitters

#### 2.1.1 On-off keying

On-off keying (OOK) is a widely adapted version of binary ASK. With OOK, the bits 1 and 0 are represented by the presence and absence of the carrier. Examples of OOK-modulated signals are depicted in Fig. 2.1. The modulator is clocked with a baseband clock whose frequency is \(f_{BB}\) and the transmit bit is toggled once per \(T_{BB}\). In this scenario, bit 1 is conveyed by the presence of the carrier and bit 0 by its absence. In the upper waveform, the phase is maintained all the time throughout transmission. This enables coherent reception. In the lower one, the phase becomes scrambled when the carrier is not generated. This generally prevents coherent reception.

One key feature of OOK in terms of power consumption is that the RF signal is output only when bit 1 is transmitted. Thus, the power amplifier (PA) needs to be active ideally only 50% of the time if bits 0 and 1 are equiprobable. This can enable low power consumption. However, low power consumption does not directly signify good energy efficiency. Instead, some choices that decrease power consumption may deteriorate error performance. For OOK to realize the same error performance with BPSK, the SNR must be greater. In other words, if an OOK transmitter outputs the same amount of peak output power as a BPSK

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\(^1\)This statement is based on the author’s observations from e.g. the narrowband ULP transmitter publications in the references of this thesis and furthermore other works cited by them.
transmitter and the output signals of both transmitters require equal noise bandwidths, the OOK transmitter possibly consumes less power but its uplink performance is also inferior.

One benefit of OOK is that it can be received noncoherently. In other words, carrier synchronization [5] is not required between the transmitter and receiver. For example, a noncoherent envelope-detecting receiver [7] can be utilized to receive the data. If noncoherent reception is utilized, the carrier phase loses its significance. This means in practice that the local oscillator (LO) of the transmitter can be switched off when zeros are transmitted. The LO and PA consume significant power in transmitters and an OOK transmitter can be implemented without other high-power blocks besides them [10]. If both of these blocks are switched off during the transmission of zeros, a low overall transmitter power consumption can be achieved. Toggling the LO off and back on generally scrambles the carrier phase as is depicted in the lower waveform in Fig. 2.1. Thus, it may be challenging to exploit the LO duty-cycling when coherent OOK transmission is desired. ULP transmitters typically have low output powers and the LO generation consumes significant power compared to the PA. To conserve power, the LO is switched off during the transmission of zeros in most published ULP OOK transmitters. This generally scrambles the phase which prevents coherent reception. Hence, the discussion related to OOK in this thesis focuses on noncoherent OOK.

One drawback of OOK is that, to achieve a given BER, coherent OOK requires approximately 3 dB more SNR per bit than BPSK and noncoherent OOK slightly more\(^2\). Thus, assuming equal noise bandwidths, a BPSK transmitter can achieve the same uplink range as an OOK transmitter with lower average output power. In such a case, use of BPSK enables lower PA power consumption. However, in ULP transmitters, the transmitted power and therefore PA power consumption may be low. Consequently, the power consumption may be dominated by the LO generation instead of power amplification [11]. In such a case, the SNR requirement has less significant meaning in terms of energy efficiency and the best modulation scheme can be the one that requires the least active time per bit – the less time the LO and PA are active, the less energy is consumed. This shall be discussed more in Chapter 4. Another drawback of OOK is that, due to the non-constant envelope of the signal, the power spectral density (PSD) of the signal sidelobes is greater compared to, for example, continuous-phase FSK. However, envelope shaping [12] can be used to improve the sidelobe attenuation.

Before discussing the details about BPSK and BFSK, we shall define a new modulation-scheme-specific factor called the *active ratio*, introduced in [I] and denoted as \(R_{t, \text{mod}}\). In the subscript, \(t\) is used to highlight the fact that the factor is related to the time domain. The abbreviation *mod* refers to the word modulation scheme and is replaced by the abbreviation of the modulation scheme in question. For instance, the active ratio of OOK is denoted as \(R_{t, \text{OOK}}\). This factor is used in the energy efficiency comparison of the modulation schemes\(^2\) This applies to the AWGN channel, discussed later in Section 2.2.
in Chapter 4. The significance of the active ratio follows from the fact that, as discussed in [I] and [II], a transmitter’s energy consumption per bit depends on the active time per bit. Regarding an OOK transmitter, the LO and PA blocks that often dominate the power consumption are ideally active for 50% of the time assuming that transmit bits 0 and 1 are equiprobable. Their active time per bit is thus $0.5 \cdot T_{BB}$. We define the active ratio as the multiplier in front of $T_{BB}$ here. Noncoherent OOK hereby has an active ratio of $R_{t,OOK} = 0.5$ [I]. It can be generalized that, when the LO and PA dominate the power consumption, a transmitter’s energy consumption per bit is roughly

$$EPB = R_{t,mod} \cdot (P_{LO} + P_{PA}) \cdot T_{BB},$$

where $P_{LO}$ and $P_{PA}$ are the active power consumptions of the LO and PA, respectively [I]. The equation assumes that the LO and PA can be switched on and off with a negligible power overhead. In summary, the active ratio is practically the fraction of $T_{BB}$ that the carrier is transmitted per bit.

### 2.1.2 Binary phase-shift keying

Fig. 2.2 shows an example of a signal modulated using BPSK. With BPSK, the bits 0 and 1 are represented by RF waveforms the length of $T_{BB}$ with the same amplitude and frequency but different phases. Antipodal waveforms with a phase difference of $180^\circ$ are generally used to minimize the error probability [5], [6]. In terms of output power, BPSK is more energy-efficient than OOK and BFSK – to achieve a given bit error probability, it requires less SNR per bit. However, as was mentioned in Section 2.1.1, the benefit of this is relative to the output power requirement and power consumption of circuitry other than the PA. We shall return to this topic in Chapter 4. As the carrier is transmitted continuously, the LO and PA are active for $T_{BB}$ per bit and the active ratio is $R_{t,BPSK} = 1$ [I].

To be able to transmit BPSK data, a transmitter must be able to toggle the phase of the RF signal. This generally requires additional circuitry that may consume some power. For example, a resonant buffer is used in [12] and a multiplexer in [13] to connect two RF carrier signals with opposite phases to a PA.
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Figure 2.3. Examples of signals modulated using binary frequency-shift keying (BFSK). The duration of a bit is $T_{BB}$. Case 1 depicts Sunde’s FSK. Case 2 depicts BFSK with a discontinuous phase.

At the receiver side, coherent BPSK detection requires a reference waveform accurate in frequency and phase [6], [7] with the received signal. Thus, carrier synchronization is generally required to minimize the bit error probability. Investing resources in the carrier synchronization, however, opens up the possibility to use other modulation schemes that can exploit the synchronization but offer higher data rate and spectral efficiency such as M-ary PSK or QAM. This may be one of the reasons why coherent BPSK has been less popular than noncoherent OOK and BFSK within published ULP narrowband transmitters. As with OOK and BFSK, coherent reception can be avoided also in the case of BPSK. This is achieved by using differential encoding which, however, leads to a minor degradation of error performance [6]. One drawback of BPSK is that, despite the constant envelope, a BPSK signal produces more out-of-band power than an OOK signal [6].

2.1.3 Binary frequency-shift keying

With BFSK, the bits 0 and 1 are represented by two carrier tones at different frequencies $f_1$ and $f_2$. Fig. 2.3 presents two forms of BFSK-modulated signals. The upper waveform depicts Sunde’s FSK [5], [6], i.e. an FSK scheme where $f_1$ and $f_2$ are both multiples of the bit rate $R_b = 1/T_{BB}$, the frequency separation between $f_1$ and $f_2$ is $\Delta f = R_b$, and phase continuity is maintained throughout the transmission even when the data bit changes from 0 to 1 or vice versa. The benefit of this phase continuity is that the side lobe PSD is decreased and thus less interference is produced outside the signal band compared to an FSK signal with a discontinuous phase, a BPSK signal or an OOK signal [5], [6]. This is a considerable advantage. The lower waveform shows an example of a BFSK signal with a discontinuous phase. With BFSK, the carrier is transmitted continuously and the active ratio is therefore $R_{t,BFSK} = 1$ [I].

BFSK can be received noncoherently and, thus, no carrier synchronization is required. While coherent BFSK reception is possible, it increases both transmitter and receiver complexity. The receiver needs to have knowledge of the frequencies and phases of both of the two signal tones [7] which implies that
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Phase-locking is required. Considering the transmitter side, to enable coherent reception, the transmitter must be able to maintain the phases of the two generated carrier tones. This may require use of two continuously running oscillators which increases power consumption. This increased complexity required by coherent BFSK only brings reduced benefit – the error performance compared to noncoherent BFSK detection is significantly better only at relatively low signal-to-noise ratios [7]. According to [4], the performance improvement of coherent BFSK over noncoherent is only 0.8 dB for bit error ratios smaller than $10^{-4}$. The limited performance improvement may not be worth the increased transmitter and receiver complexity. Instead, if coherent reception is implemented, coherent BPSK may be preferable over coherent BFSK as it requires phase coherence with a single tone only and the error performance is better compared to coherent BFSK [7]. For these reasons, noncoherent BFSK may be preferable over coherent BFSK in ULP applications. Thus, when BFSK is discussed in this thesis, noncoherent reception is assumed. For signal orthogonality, noncoherent BFSK requires that the frequency separation is $\Delta f = 1/T_{BB}$ [4]. If another frequency separation is used, BFSK might not reach the error performance of noncoherently detected binary orthogonal FSK.

2.2 Error performances of OOK, BPSK and BFSK

A radio transmitter generates a modulated signal which passes through a medium before reaching the antenna of a receiver. The receiver typically demodulates the signal and in that way attempts to reproduce the transmitted information-bearing signal, be it an analog signal, e.g. a piece of music or speech in traditional FM radio, or a digital signal in a wireless local area network (WLAN). However, the received RF signal is not an exact replica of the transmitted signal. Instead, it is affected by the antenna, transmission channel and environment, typically in a way that degrades the quality of the signal. Consequently, when the received signal is demodulated, the original information-bearing signal cannot be reproduced perfectly. In the case of digital signals, this means that, with some probability, bit errors are produced, i.e. a bit 0 is interpreted as bit 1 or vice versa.

Some nonidealities in a received signal are, for instance, noise, reflections, signals from other transmitters and frequency shift [1], [4], [5], [6], [7]. Thermal noise is added to the signal and additional noise is introduced by the receiver circuitry. The transmitted signal can be reflected from the environment which adds reflections. In a real terrestrial radio environment, a received signal also includes interfering signals from other transmitters. Frequency shift may occur due to Doppler shift – the frequency of the signal or its reflections may shift if the transmitter, receiver or nearby reflecting objects move. Different modulation schemes have differing error performance when the signal is affected by these and other phenomena. Considering a modulation scheme’s capability to cope
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with these challenges is of importance for establishing a reliable radio link.

In error performance analysis of a modulation scheme, the channel and its effect on the signal needs to be modeled. One basic channel model that is used for comparing the performance of modulation schemes is the additive white Gaussian noise (AWGN) channel [5], [6] which solely adds AWG noise to the signal and other non-idealities are omitted. Analyzing the error performance in the AWGN channel provides a view of the best achievable error performance with a modulation scheme and, in the presence of other non-idealities, the performance is generally degraded [6]. The error performances of OOK, BPSK and BFSK in the AWGN channel are well described in the literature [1], [4], [5], [6], [7].

In the AWGN channel, the bit error ratios with OOK, BPSK and BFSK are as follows. When an OOK signal is received noncoherently using a receiver with an envelope detector, the theoretical BER [7] is

\[
BER_{OOK} = \frac{1}{2} \left[ 1 - Q\left(\sqrt{2 \cdot \gamma}, b_0\right) \right] + \frac{1}{2} \exp\left( -\frac{b_0^2}{2} \right). \tag{2.3}
\]

Here \(Q(a, b)\) is the Marcum-Q function and \(\gamma = A^2/2\sigma^2\) is the SNR\(^3\) in the envelope detector output while the carrier is being received. \(A\) and \(\sigma\) are parameters related to the envelope detector output distributions that are discussed more in Section 3.2.4. The decision of whether the received bit is 0 or 1 is made based on the amplitude of the envelope detector output. The amplitude is compared with the threshold \(b_0\) to make the decision. The optimal threshold level depends on \(\gamma\) and is ideally set to a value that statistically minimizes the probability of a bit error. The optimal threshold at a given \(\gamma\) is well approximated by \(b_0 = \sqrt{2 + \gamma/2}\) [7], [14]. The mathematics behind (2.3) is related to the distributions of the envelope detector output. When the carrier is not received, the envelope detector output is affected by noise only in which case it is Rayleigh-distributed [4] and affected by \(\sigma\) only. When the carrier is received, the output is Rician-distributed [4] and it depends on both \(A\) and \(\sigma\). Fig. 2.4 shows examples of Rayleigh and Rician distributions with \(A = 1\) and \(\sigma = 0.2235\), corresponding to \(\gamma\) of 10 dB. The derivation of (2.3) is given e.g. in [7].

When error performances are compared, the BERs are typically plotted against the SNR per bit [4], [6], often denoted as \(\gamma_b\) or \(E_b/N_0\).\(^4\) \(E_b\) is the signal energy per bit and \(N_0\) is the noise power spectral density. In this work, we discuss radio communication and generally consider the noise as limited by thermal noise. The PSD of thermal noise is given by \(k \cdot T\), where \(k\) is the Boltzmann constant and \(T\) is the noise temperature [4]. For this, we define the noise PSD as \(N_0 = k \cdot T\). Note that it can be deduced that the SNR per bit \(E_b/N_0\) is unitless because the units of \(E_b\) and \(N_0\) are joules and watts per hertz, respectively.

\(^3\)In this work, the symbol \(\gamma\) refers to the SNR in general, not only the SNR in an envelope detector output.

\(^4\)Instead of the term SNR per bit, some sources may also use the term bit energy to noise spectral density ratio for \(\gamma_b\) and \(E_b/N_0\) [6].
In the case of OOK, the BER as a function of the SNR per bit is obtained as follows. Since the carrier is generated on average only 50% of the time, the SNR per bit is obtained from \( \gamma \) by dividing it by two \([7]\), i.e. as \( E_b/N_0 = \gamma/2 \). Conversely, it can be written that \( \gamma = 2E_b/N_0 \). Substituting this to (2.3) gives

\[
BER_{OOK} = \frac{1}{2} \left[ 1 - Q\left( \sqrt{4 \frac{E_b}{N_0}}, b'_0 \right) \right] + \frac{1}{2} \exp\left( - \frac{b'_0^2}{2} \right). \tag{2.4}
\]

With the BER written as a function of \( E_b/N_0 \), it is to be noted that the approximate optimum threshold is \( b'_0 = \sqrt{2+E_b/N_0} \).

The theoretical BER with coherent BPSK \([5], [7]\) is

\[
BER_{BPSK} = \frac{1}{2} \cdot \text{erfc}\left( \sqrt{\frac{E_b}{N_0}} \right). \tag{2.5}
\]

Here, \( \text{erfc}(x) \) is the complementary error function. The theoretical BER with noncoherent BFSK \([4], [5], [7]\) is

\[
BER_{BFSK} = \frac{1}{2} \cdot \exp\left( - \frac{1}{2} \frac{E_b}{N_0} \right). \tag{2.6}
\]

Note that (2.6) only applies if the two BFSK transmit waveforms are orthogonal.
The BERs as obtained with (2.4), (2.5) and (2.6) are plotted in Fig. 2.5 against the signal-to-noise ratio per bit. At decent BERs below $10^{-3}$, the difference between the performance of noncoherent OOK and BFSK is only minor, in the range of 0.5 dB or less. For BER $= 10^{-5}$, coherent BPSK requires 9.6 dB of SNR per bit while noncoherent OOK and BFSK require 13.1 and 13.4 dB, respectively, i.e. 3.5 and 3.8 dB more than BPSK. It is to be noted that these BER curves assume ideal signals and ideal detection. In reality, due to the nonidealities of radio transmitters, the generated output signals are not ideal and the SNR required for a given BER may be higher than the theoretical value. The receiver circuitry affects the error performance as well.

A low required signal-to-noise ratio per bit implies that the modulation scheme requires low average transmitted power to achieve a certain error probability. Hence, looking at Fig. 2.5, BPSK seems to be more optimal a choice for a transmitter than OOK or BFSK. Regarding transmitters in which the PA dominates the power consumption, an output power decrement of 3.5 to 3.8 dB indeed translates to significant power saving because it implies that PA power consumption decreases by roughly 55% to 58%. However, in ULP transmitters, the output power and PA power consumption can be low. Instead of the PA, other blocks such as the LO may consume a significant proportion of the total power or even dominate the power consumption. In such a case, this kind of a PA power consumption decrement does not necessarily lead to the lowest energy consumption per bit. Instead, it could be more beneficial to use modulation schemes such as OOK, PPM or DPPM that enable a lower active time per bit of the LO and PA. This is suggested by the modulation energy efficiency comparison in Chapter 4 which also accounts for the active ratios of the compared modulation schemes. Regarding the comparison, it is more convenient to express the error performances as a function of $\gamma$ as opposed to SNR per bit. With BPSK and BFSK, the SNR $\gamma$ and SNR per bit $\gamma_b$ (i.e. $E_b/N_0$) are equal. Thus, the BERs with BPSK and BFSK are, as given in [7], simply

$$BER_{BPSK} = \frac{1}{2} \cdot \text{erfc}(\sqrt{\gamma})$$

(2.7)

and

$$BER_{BFSK} = \frac{1}{2} \cdot \exp\left(-\frac{\gamma}{2}\right).$$

(2.8)

Note that signal-to-noise ratio $\gamma$ does not refer to the ratio between average signal power and noise power. Instead, according to a waveform-level PER simulation study discussed in Section 3.3, it refers to the ratio

$$\gamma = \frac{E}{E_n},$$

(2.9)

where $E$ is the energy of the carrier wave over the time interval $T_{BB}$ and $E_n$ is the energy of the noise in the noise bandwidth over the time interval $T_{BB}$. The average-signal-power-to-noise-power ratio is used in some calculations and it is
denoted in this work as
\[ \gamma_{avg} = \frac{P_{\text{sig}}}{P_n}, \]  
(2.10)
where \( P_{\text{sig}} \) and \( P_n \) are the average signal power and the noise power in the noise bandwidth, respectively. The corresponding average-signal-power-to-noise-power ratio required by a modulation scheme is denoted as \( \gamma_{\text{avg,req}} \).

### 2.3 About energy efficiency and energy consumption per bit

In the vast majority of ULP transmitter publications, the energy efficiency has not been evaluated using proper metrics. In several publications, the EPB has been used as the measure of the energy efficiency. The EPB can be written [10] as
\[ \text{EPB} = \frac{P_{TX}}{R_b}, \]  
(2.11)
where \( P_{TX} \) and \( R_b \) are the total power consumption and data rate of a transmitter, respectively. The EPB is a decent metric for the energy consumption of a transmitter as it expresses how much energy is consumed per transmission of a bit. The lower the EPB, the less energy is consumed for transmitting a given amount of data and, conversely, the more data can be transmitted with a given amount of available energy. Nonetheless, it does not enable a fair direct comparison between transmitters [10]. It only considers the consumed energy and not at all what is achieved with it. It completely neglects various important metrics such as the output power, noise bandwidth and utilized modulation scheme [10], [I]. Considering the comparison of transmitters, the major drawback of using EPB as a figure of merit for energy efficiency is that a low EPB can often be achieved by use of a high data rate. A high data rate, however, implies a wider noise bandwidth, greater noise power and degraded uplink performance as is explained next.

With OOK, BPSK and BFSK, a modulator toggles the carrier amplitude, phase or frequency at a baseband clock frequency \( f_{BB} = 1/T_{BB} \). The data rate \( R_b \) is simply increased by increasing \( f_{BB} \) and, as per (2.11), this reduces the EPB. However, as discussed in [III], a drawback follows in the frequency domain as signal bandwidth is directly proportional to \( f_{BB} \). This is explained by the dilation property of Fourier transform [1]. If a signal is compressed in the time domain by the dilation factor \( a \), its spectrum expands by factor \( a \) as well. Considering OOK, BPSK and BFSK transmitters, if \( f_{BB} \) is scaled up by the factor \( a \) for a greater data rate, \( T_{BB} \) is compressed by \( a \) and the signal power is spread over a wider band. Thus, the power spectral density of the signal decreases by the factor \( a \). As said in [III], this assumes that, as \( T_{BB} \) is scaled down, the whole baseband signal with any envelope shaping is compressed which also maintains
Figure 2.6. An example of the expansion of signal bandwidth as data rate is increased in the case of noncoherent OOK. (top left) OOK signal at 1x data rate, (top right) OOK signal at 10x data rate, (middle) signal spectra, and (bottom) signal spectra normalized to the data rate $R_b$.

spectral shaping. The consequence of the higher data rate is thus a wider signal bandwidth, a wider noise bandwidth and increased noise power [10], [I], [II].

Fig. 2.6 shows a simplified depiction of the problem with high data rates in the case of a noncoherent OOK signal. The subfigures in the top left and top right show clips of OOK-modulated signals with data rates of 1 bps and 10 bps, respectively. The signal amplitudes are equal and therefore the peak and average powers as well. The carrier sine wave cycles are not distinguishable because the carrier frequency is much higher than the bit rates. The subfigure in the middle row shows the signal spectra. It can be seen that the spectrum is expanded tenfold with the tenfold data rate. The bottom subfigure shows the spectra normalized to the bit rate $R_b$. The PSD level is 10 dB lower with the 10x data rate. To receive the signal with the 10x data rate, a receiver must use a tenfold bandwidth and the noise power in the band is tenfold. Hereby, as the signal powers with both data rates are equal, the maximum achievable SNR is 90% lower in the case of the 10x data rate and the uplink performance is inferior.

The connection of the above issue to the insufficiency of EPB as a figure of merit for energy efficiency is that the signals of Fig. 2.6 could be generated by two transmitters whose LO, PA and total power consumptions are identical.
The one with the higher data rate would have a 90% lower EPB because of the shorter time required per transmission of a bit but the signal SNR would be 10 dB lower as well. The lower SNR would not be reflected in the EPB at all. The transmitter with the lower data rate may consume more energy per bit but is also expected to achieve a higher uplink range. Due to this, the EPB is not a good metric for the energy efficiency of a transmitter. Looking at the time-domain signals in Fig. 2.6, one observation can be made: with the 10x data rate, \( T_{BB} \) is 90% lower and, consequently, the transmitted energy per bit is 90% lower. Adjusting the data rate through \( f_{BB} \) thus scales the consumed energy per bit and the transmitted energy per bit ideally by the same factor. Based on this, it can be deduced that the adjustment of \( f_{BB} \), in reality, has no impact on energy efficiency. This point of view was briefly mentioned in [I].

The spectra of Fig. 2.6 were produced by generating 512 time-domain signal vectors the length of 100 seconds with each bit rate and by calculating the average spectrum for a smoother plot. The signal sample rate and carrier frequency were 20 kHz and 2 kHz, respectively. To model noncoherent OOK, the carrier phase was randomized for each sequence of ones, i.e. every time the bit changes from zero to one. This explains why no distinct peak is present in the spectrum at the carrier frequency.

Adjusting only the data rate through \( f_{BB} \) is hereby not expected to affect the energy efficiency of a transmitter. However, as per [II], it may be possible to improve the energy efficiency if the output power is increased by the same factor with \( f_{BB} \) to maintain the SNR. In ULP transmitters, most power is generally consumed in carrier generation and power amplification [10]. Moreover, with the low output powers of ULP transmitters, the carrier generation tends to consume a significant share of the total power and may even dominate the power consumption. With a high \( f_{BB} \), \( T_{BB} \) is low and the LO consumes less energy per bit. Thus, if \( f_{BB} \) is scaled up and the output power is scaled accordingly to maintain the SNR, the energy consumption of the LO per bit reduces, the global efficiency approaches the efficiency of the PA, and the overall energy efficiency could improve. However, this assumes that the output power can be scaled up without increasing the LO power consumption significantly.

As a case example, in the style of [II], let us consider what kind of an effect a tenfold increase of data rate has on the EPB of a transmitter whose data rate and output power are \( R_b = 100 \) kbps and \( P_{out} = -20 \) dBm = 10 \( \mu \)W, respectively. The transmitter contains an LO that consumes \( P_{LO} = 100 \) \( \mu \)W of power and the drain efficiency of the PA is \( \eta_{PA} = 40\% \). Thus, the PA power consumption is \( P_{PA} = 25 \mu \)W. Assuming that the LO and PA dominate the power consumption, the total EPB of the transmitter is approximately

\[
EPB = \frac{P_{LO} + P_{PA}}{R_b} = \frac{100 \mu \text{W} + 25 \mu \text{W}}{100 \text{ kbps}} = 1.25 \text{ nJ/bit.} \tag{2.12}
\]

Let us then consider how the EPB of the example transmitter changes if the data rate is increased tenfold to \( R_{b2} = 1 \) Mbps. Due to the dilation property, the
signal bandwidth scales linearly with $f_{BB}$. Therefore, the output power is scaled linearly with $f_{BB}$ to maintain the SNR and, ideally, the uplink range. At the data rate of 1 Mbps, the output power must be tenfold i.e. $P_{out2} = -10$ dBm = 100 $\mu$W. Assuming that the PA drain efficiency remains unchanged, the corresponding PA power consumption is $P_{PA2} = 250$ $\mu$W. In this configuration, the EPB is

$$EPB = \frac{P_{LO} + P_{PA2}}{R_{b2}} = \frac{100 \mu W + 250 \mu W}{1 \text{ Mbps}} = 0.35 \text{ nJ/bit.} \quad (2.13)$$

The total EPBs with the lower and higher data rate are thus 1.25 and 0.35 nJ/bit, respectively, i.e. increasing the data rate improved the EPB by 72%. As the output power was scaled to maintain the SNR, the transmitter would ideally achieve the same uplink performance with the higher data rate. Hereby, in this example case, the use of the higher data rate results in a higher energy efficiency. At the lower data rate, the power consumption is dominated by the LO which results in a lower global efficiency. The output power of 10 $\mu$W in this configuration is relatively low compared to the total power consumption of 125 $\mu$W and the global efficiency is 8%. At the higher data rate, the power consumptions of the LO and PA per bit are more balanced and the global efficiency is improved – it is $\eta = P_{out2}/(P_{LO} + P_{PA2}) \approx 28.6\%$.

The point of this example case was to bring up the fact that it may conserve energy on the transmitter side if a high data rate is used. However, it is to be kept in mind that the PA output power scaling needs to be taken into account when different data rates are considered. The higher the data rate is, the more output power is required to achieve a given SNR. Despite the possible benefits, use of a high data rate can only be exploited limitedly as, typically, the spectrum of the transmitted signal is limited by a radio standard and only a limited amount of bandwidth is available [10], [II].

This section explained the issue of using EPB as an FOM for energy efficiency and in general for comparing transmitters. The main problem is that EPB neglects various important metrics such as output power, noise bandwidth and the SNR required by the utilized modulation scheme which together affect the uplink performance. In the next section, a figure of merit is derived that takes into account these metrics and avoids improper data rate dependency.

### 2.4 Figure of merit for energy efficiency

Before reviewing previously published transmitters, let us review the derivation of the transmitter energy efficiency FOM presented in [I]. Compared to many other energy FOMs which have been used in ULP transmitter publications, the FOM of [I] enables a more rational comparison between transmitters that may differ in terms of modulation, consumed power, output power and data rate. As was discussed in Section 2.3, the EPB neglects various important metrics and is not a fair and suitable FOM for energy efficiency. Another FOM has been
discussed, for instance, in [10] which is

\[ FOM_{E_{\text{nom}}} = \frac{EPB}{P_{\text{rad,out}}} \]  

(2.14)

where \( P_{\text{rad,out}} \) is the radiated output power. Using (2.11), the equation can also be written as \( FOM_{E_{\text{nom}}} = \frac{P_{TX}}{(R_b \cdot P_{\text{rad,out}})} \). Thus, (2.14) takes into account the output power and also reflects power efficiency which is an improvement compared to the EPB. However, in a similar way to the EPB, \( FOM_{E_{\text{nom}}} \) is highly dependent on data rate \( R_b \) and favors high-data-rate transmitters whose output signals may suffer from a low SNR due to a wide bandwidth and the consequent increased amount of noise [I]. Moreover, it neglects the effect of modulation although, as discussed in [I], the SNR requirement of the utilized modulation scheme impacts the achievable uplink performance. The SNR requirement is a factor in receiver sensitivity – the less SNR is required to demodulate the signal with an acceptable error probability, the better is the receiver sensitivity which can be achieved [I]. Thus, a lower SNR requirement, in part, enables a longer uplink range. Furthermore, another FOM is \( \text{EPB}/\eta \) [15], where \( \eta \) is the power efficiency. In a similar fashion to \( FOM_{E_{\text{nom}}} \), this FOM is directly proportional to the EPB and suffers from similar drawbacks as \( FOM_{E_{\text{nom}}} \) [I].

In [16], it is discussed that also the communication range should be considered in an FOM. This is a significant improvement compared to the above FOMs. However, the FOM presented in [16] applies to transceivers and does not directly apply to transmitter-only systems. The FOM provided in [16] is

\[ \text{Energy Efficiency} = LS \cdot \frac{R_b}{P_{TX} + P_{RX}} \]  

(2.15)

where \( R_b \) is the data rate and \( P_{TX} \) and \( P_{RX} \) are the transmitter and receiver power consumption, respectively. \( \frac{R_b}{(P_{TX} + P_{RX})} \) is effectively the reciprocal of the combined energy consumed by a transmitter and a receiver per bit. \( LS \) is the link strength [16]. It is defined as

\[ LS = \frac{P_{\text{out}}}{S_{RX}} \]  

(2.16)

where \( P_{\text{out}} \) and \( S_{RX} \) are the transmitter’s output power and receiver sensitivity, respectively. The factor \( LS \) determines how much the output signal may attenuate before it decreases to the power level at which the receiver sensitivity is defined. This measure of allowed attenuation is highly related to the achievable uplink range. According to (2.15), a good energy efficiency is achieved if the \( LS \) is high, i.e. if the transceiver can achieve a high link range, with respect to the energy consumed per bit. The unit of the result of (2.15) is bit/J and the higher the FOM of a transceiver is, the more energy-efficient it is. The FOM accounts for bandwidth and modulation through the receiver sensitivity \( S_{RX} \). The wider the bandwidth, the higher \( S_{RX} \) is and the more output power is required for a given link strength. The less SNR the modulation scheme requires for the targeted error probability, the lower \( S_{RX} \) is and the less output power suffices.
As discussed in [I], (2.15) can be modified as follows to make it applicable to transmitter-only systems instead of transceivers. Firstly, we write the new energy efficiency FOM as the reciprocal of (2.15) and remove $P_{RX}$ to consider a transmitter only as opposed to a transceiver. The FOM becomes

$$FOM = \frac{P_{TX}}{LS \cdot R_b}. \quad (2.17)$$

Furthermore, with (2.11), (2.17) can be written as

$$FOM = \frac{EPB}{LS}, \quad (2.18)$$

where EPB is now the EPB of a transmitter only [I]. Secondly, we redefine the link strength, $LS$. The $LS$ in (2.16) considers how much the output power may attenuate before reaching the sensitivity, $S_{RX}$, i.e. the input power level that the receiver in the transceiver system requires for the specified error probability. In a transmitter-only system, no such reference receiver exists whose sensitivity could be used to define the $LS$. For this, the $LS$ was specified in [I] as the ratio of the maximum SNR achievable with the output signal, $\gamma_{max}$, to the SNR that the utilized modulation scheme requires for the desired error probability, denoted as $\gamma_{req}$ [I]. The link strength is therefore redefined as

$$LS' = \frac{\gamma_{max}}{\gamma_{req}}. \quad (2.19)$$

Replacing the $LS$ in (2.18) with $LS'$, we get a short form of the transmitter energy efficiency FOM,

$$FOM = \frac{EPB \cdot \gamma_{req}}{\gamma_{max}}, \quad (2.20)$$

as given in [I]. This can also be written in decibels as

$$FOM = 10 \cdot \log_{10} \frac{EPB}{1\text{ J/bit}} + (\gamma_{req})_{\text{dB}} - (\gamma_{max})_{\text{dB}}. \quad (2.21)$$

The maximum SNR achievable with the output signal is calculated as

$$\gamma_{max} = \frac{P_{out}}{P_n}, \quad (2.22)$$

where $P_{out}$ is the transmitter’s output power and $P_n$ the noise power in the noise bandwidth, given by

$$P_n = k \cdot T \cdot BW_N. \quad (2.23)$$

Here $k$, $T$ and $BW_N$ are the Boltzmann constant, noise temperature in Kelvin and receiver noise bandwidth, respectively [4], [17]. In this work, noise temperature $T = 298$ K is assumed. Referring to ASK, PSK and FSK, it is stated in [17] that noise bandwidth is equal to the symbol rate. In the case of OOK, BPSK

\footnote{Calculating the reciprocal in [I] was purely just the author’s preference for having the unit in the style of the EPB as J/bit or dBJ/bit in decibels.}
and BFSK, the symbol rate is equal to the data rate and data rate $R_b$ may thus be used as the noise bandwidth $BW_N$.\footnote{For instance, if an OOK, BPSK or BFSK signal is received at data rate $R_b = 1$ Mbit/s, the receiver noise bandwidth is ideally $BW_N = R_b = 1$ MHz. Noise bandwidth is discussed more related to receiver simulations in Section 3.3.} Substituting (2.23) to (2.22) gives the maximum SNR achievable with the output signal as

$$\gamma_{max} = \frac{P_{out}}{k \cdot T \cdot BW_N}. \quad (2.24)$$

In this work, the FOM of transmitter circuitry is considered. For that, the $\gamma_{max}$ here neglects any power gain that could be achieved, for example, with antenna gain. An extended form of the FOM is obtained by substituting (2.11) and (2.24) to (2.20) which, as per [I], yields

$$FOM = \frac{P_{TX} \cdot k \cdot T \cdot BW_N \cdot \gamma_{req}}{R_b \cdot P_{out}}. \quad (2.25)$$

The unit of this FOM is J/bit or dBJ/bit in decibels. The lower the FOM of a transmitter is, the more energy-efficient it is. This FOM accounts for the noise bandwidth, SNR and the SNR requirement of the modulation scheme, and, in this way, it is more comprehensive than the FOMs which have been used in the majority of earlier ULP transmitter publications. The noise bandwidth should be accounted for in an FOM because it impacts the SNR and, therefore, the uplink performance. The FOM favors transmitters whose power efficiency is high which is one indicator of good energy efficiency. The greater the power efficiency, the greater $P_{out}$ and $\gamma_{max}$ are achieved with respect to the consumed power and EPB. As was discussed in Section 2.3 and [I], adjusting the data rate through $f_{BB}$ ideally has no effect on the energy efficiency. Correspondingly, scaling $f_{BB}$ has no major impact on this FOM. Scaling $f_{BB}$ scales both the data rate, $R_b$, and the noise bandwidth, $BW_N$, equally. It can be seen in (2.25) that this ideally does not affect the FOM. The differing SNR requirements between modulation schemes are taken into account in $\gamma_{req}$. Accounting for the modulation scheme and its SNR requirement is reasonable – after all, the best achievable uplink performance does not depend only on the maximum SNR achievable with the output signal but rather on this maximum SNR relative to the SNR required by the modulation scheme.

In the FOM calculation, the utilized numerical values must be chosen properly. In (2.25), $P_{TX}/R_b$ represents the EPB and, thus, $P_{TX}$ is the transmitter power consumption when modulated data is transmitted at the data rate $R_b$. $BW_N$ is the noise bandwidth required by the output signal when the data rate is $R_b$. $P_{out}(k \cdot T \cdot BW_N)$ is the maximum SNR achievable with the output signal and, because $\gamma_{req}$ is used, $P_{out}$ is the peak output power in the case of OOK, BPSK, BFSK, PPM and DPPM. Alternatively, as per [I], the average output power $\bar{P}_{out}$ can be used in the calculation in which case $\gamma_{req}$ must be replaced with the average-signal-power-to-noise-power ratio required by the modulation scheme,
denoted in this work as $\gamma_{avg,req}$ [I]. This way the FOM can be rewritten as

$$FOM = \frac{P_{TX} \cdot k \cdot T \cdot BW_N \cdot \gamma_{avg,req}}{R_b \cdot P_{out}}. \quad (2.26)$$

Each of (2.20), (2.25) and (2.26) is expected to give the same result when the calculation is done properly. Examples of the FOM usage and calculations are provided in [I] and Sections 2.5.5, 5.1.5, 5.2.5 and 5.3 of this thesis.

Note that the FOM enables comparison of various types of transmitters. Most transmitter publications provide sufficient information for calculating the FOM [I]. Power consumption, output power and data rate are generally reported in publications. Noise bandwidth $BW_N$ is often known, or can be deduced, based on the utilized modulation scheme and the data rate. Furthermore, the SNR requirement depends on the modulation scheme. As pointed out in [I], also the FOMs of ultra-wide band transmitters can be assessed. However, if a UWB transmitter uses duty-cycling, the noise bandwidth is determined rather by the UWB pulse duration as opposed to data rate or baseband clock frequency. It is to be noted that transmitters should be compared considering a common error performance target. In this work, we consider a BER target of $10^{-5}$ or an equivalent PER target with 48-bit packets as the transmitters reported in Sections 5.1 and 5.2 transmit DPPM data in packet-mode. When comparing transmitters, it is reasonable to also compare the link strengths. Using the peak output power, the link strength can be written as

$$LS = \frac{P_{out}}{k \cdot T \cdot BW_N \cdot \gamma_{req}}. \quad (2.27)$$

Correspondingly, the link strength using the average output power is

$$LS = \frac{P_{out}}{k \cdot T \cdot BW_N \cdot \gamma_{avg,req}}. \quad (2.28)$$

In a similar fashion to the FOM, the link strength can be calculated using different methods but each of them is expected to yield the same result.

### 2.5 Review of state-of-the-art ULP transmitters

This section discusses state-of-the-art ULP narrowband transmitters with submW power consumption. The discussion is limited to active transmitters that generate the LO signal autonomously. Some transmitters use a received reference RF signal in LO generation [18] or backscattering techniques [19] which can enable reduction in TX power consumption. However, they typically require a relatively strong reference signal which limits their use. These types of works are beyond the scope of the present discussion but, nonetheless, it is good to know that such methods could enable lower power consumption.
2.5.1 Overview

Within published transmitters, the lowest power consumptions have generally been achieved using simple binary modulation schemes such as noncoherent OOK or noncoherent BFSK. OOK and BFSK signals can be generated using low-complexity transmitter architectures that achieve low power through a reduced amount of circuit blocks and particularly high-power blocks [10].

Fig. 2.7 presents two popular low-complexity architectures: a direct-modulation architecture and an architecture based on a power oscillator (PO) [10]. A direct-modulation transmitter requires only three major blocks: a modulator, a carrier synthesizer, and a power amplifier, as depicted in Fig. 2.7(a). OOK is performed by switching the PA on and off in accordance with the binary data. A lower power consumption is achieved if the LO is duty-cycled as well. BFSK can be performed, for instance, by using an LC oscillator as the LO and by toggling its frequency by switching the capacitance of the LC tank. To implement BPSK, circuitry for shifting the phase [12], [13] can be inserted between the LO and PA. In the PO-based topology, shown in Fig. 2.7(b), an off-chip or on-chip metal strip acts both as the inductor of the LC tank of the oscillator and as the radiative element, i.e. the antenna. Thus, the PO directly delivers the signal to the antenna and a PA is not utilized. OOK and BFSK can be performed by toggling the amplitude and frequency of the PO, respectively.

To achieve low power and high efficiency with these low-complexity transmitters, the carrier signal must be generated with low power and power amplification must be performed efficiently. The modulator is typically clocked at the baseband frequency and only consumes minor power. Thus, power-efficient design requires a particular focus on efficient carrier signal generation and power amplification. Some popular oscillator and PA-related blocks are discussed briefly in the next two sections. A more thorough discussion on different types of carrier synthesis methods and PAs has been presented, for example, by D.-G. Lee et al. in [10]. Many published ULP transmitters have been implemented by combining the following types of LO and PA designs in various ways. A selection of publications is reviewed in Section 2.5.4 to give an overview of some ULP transmitter implementations. These transmitters and their energy efficiency FOMs are compared in Section 2.5.5.
2.5.2 ULP carrier generation methods

Popular methods [10] for generating the carrier are the use of a digitally controlled oscillator (DCO), a voltage controlled oscillator (VCO) and an oscillator whose frequency is set using a high-Q resonator. A particularly effective way to achieve a low LO power consumption has been to use a low-frequency LO (LF-LO) together with edge-combining or a frequency-multiplying PA [10]. A significant advantage of VCOs and DCOs is that they can be used to generate a wide range of carrier frequencies which enables transmission in multiple bands or sub-bands. A free-running DCO or VCO enables low complexity, a fast start-up, and low active and standby power consumption. However, frequency variation may occur due to, for example, temperature variation and supply voltage inaccuracy. Additionally, if the PO-based topology is used and an off-chip metal strip (e.g. an on-PCB copper loop) acts as the inductor of the LC tank of the oscillator, the frequency is susceptible to strong external interferers and, in addition, nearby physical objects. This has been addressed, for instance, in [20] and (III).

To avoid the frequency inaccuracy of a free-running DCO or VCO, the oscillator can be used in a phase-locked loop (PLL) or a frequency-correcting loop. This, however, generally requires additional blocks that increase power consumption to some extent. For instance, a PLL can contain a phase frequency detector, frequency divider and a charge pump [21]. However, some low-power implementations can be found in the literature. For example, in [22], a 72-μW PLL with a closed-loop bandwidth of 80 kHz is used in a 400–433 MHz 80-kbps BFSK transmitter. If the power consumption of a PLL is considered too high, its power consumption may be reduced by duty-cycling the operation of the feedback path [10]. The locking can be performed, for instance, only prior to data transmission [10], [21] after which the loop can be opened. The shortcoming of the duty-cycling approach is that the frequency may drift until the next frequency calibration.

Another method for synthesizing an accurate carrier frequency is to utilize an RF oscillator whose frequency is set using a high-Q resonator such as a film bulk acoustic resonator (FBAR) or a surface acoustic wave (SAW) resonator [10]. This avoids the power overhead of a PLL but, compared to a free-running VCO or DCO, the startup time of the oscillator may be longer and the frequency tuning range lower. Limited frequency tunability limits band and channel selection capabilities. However, more bands and channels become available if multiple resonators are used as is done in [12]. If FSK is utilized, the limited frequency tunability also limits the data rate as the frequency separation must be $\Delta f = R_b$ with noncoherent FSK for orthogonality [4]. Use of a lower frequency separation is of course possible but may degrade the error performance.

Furthermore, an accurate carrier frequency can be synthesized without a PLL using a low-power low-frequency multi-phase reference together with a

\[^{7}\text{To be more specific, the FSK tones must be multiples of } R_b \text{ for orthogonality [4].}\]
frequency-multiplying edge-combiner PA (ECPA) [10]. Such transmitter architecture was presented by J. Pandey et al. in [11] and is depicted in Fig. 2.8(a). A low power consumption is enabled as the low-frequency LO operates at a lower frequency and only the ECPA at the carrier frequency. In [11], the LO operates at 44.5 MHz and the frequency is set using a crystal oscillator (XO) that provides an accurate frequency with low phase noise. Nine LO signal phases $A_1$–$A_9$ are generated using cascaded injection-locked ring oscillators (ILROs), shown in Fig. 2.8(b). The phases, depicted in Fig. 2.9(a), are fed to the ECPA, shown in Fig. 2.9(b). The ECPA multiplies the LO frequency by a factor of 9 and, thus, the input voltage of the antenna is a sinusoidal voltage at a frequency of $9 \times 44.5 \text{ MHz} = 400.5 \text{ MHz}$. Using this method, only 24 $\mu$W is consumed by carrier generation and modulation in [11]. Also here the disadvantage is that the limited frequency tuning range of the resonator can limit the channel/band selectivity and FSK data rate.

The combination of an LF-LO and an ECPA has been utilized effectively at frequencies in the range from 400 to 427 MHz [11], [13], [23], [24]. However, the multiple phases fed to the ECPAs have been generated using ring oscillators (ROs). As the power consumption of an RO is directly proportional to frequency [24, Eq. 1], the LO power consumption with this approach can be expected to be greater at higher carrier frequencies. It is worth mentioning that the edge-combining technique has also been used in a BLE transmitter in [25] to generate a 2.4-GHz carrier. There, the circuitry related to carrier generation consumes 379 $\mu$W [25, Fig. 16] which is significantly greater than the 24 $\mu$W consumed
in [11]. However, this greater consumption may also be partly related to the stricter phase noise requirements in BLE communication.

### 2.5.3 ULP power amplifiers

Published and measured ULP power amplifiers have achieved decent PA efficiencies exceeding 40% with low output powers down to –20 dBm. At low output powers, the amplitudes over a 50-Ω load are fairly low. For example, for –20-dBm and –10-dBm power, the peak-to-peak voltages over a 50-Ω load are 63 and 200 mV, respectively. Due to such low voltages, several works such as [12] and [26] use a lowered PA supply voltage to improve the drain efficiency (DE).

Good PA efficiencies have been achieved, for example, using the ECPA in [11], a class-C PA in [22], a class-F PA in [26], an inverter-based push-pull PA topology in [12], and a switched-capacitor digital PA (SCDPA) in [25]. Some PA efficiencies reported in these works are plotted in Fig. 2.10.

Pandey et al. [11] use the ECPA of Fig. 2.9(b). The supply of the ECPA is 1.0 V and a DE of 30% is achieved at an output power of –17 dBm. According to [11, Fig. 19], if a lower output power of –23 dBm is utilized, the DE is still 27%. Natarajan et al. [22] use two parallel pseudodifferential class-C PAs with separate matching networks for two output power modes: a low-power (LP) and high-power (HP) mode. The LP-mode PA is optimized for –12-dBm output power and the HP-mode PA for –2-dBm output power. A single PA without a matching network is shown in Fig. 2.11(a). Drain efficiencies of 33%, 43% and 47% are achieved at output powers of –16, –12 and –2 dBm, respectively. At output power of –30 dBm, the DE is 10%. Based on [22, Fig. 5], the supply voltage is 0.7 V. Iguchi et al. [26] use a class-F PA, depicted in Fig. 2.11(b). Using a 0.2-V PA supply, the DE is 42% at an output power of –20 dBm.

Paidimarri et al. [12] use a push-pull PA, shown in Fig. 2.12(a). They achieve a DE of 44.4% with an output power of –9.5 dBm using a 0.5-V PA supply. In a similar fashion with Paidimarri, Chen et al. [25] also use a complementary PMOS-NMOS PA. With their SCDPA, shown in Fig. 2.12(b), a maximum DE of 41% is achieved at around –7.1-dBm output power with a supply of 0.7 V. With a
Binary modulation schemes, literature review and new FOM for transmitter energy efficiency

Figure 2.11. (a) Pseudodifferential class-C PA used in [22], and (b) class-F PA used in [26].

Figure 2.12. (a) Push-pull PA from [12]. © 2013 IEEE. (b) Switched-capacitor digital power amplifier (SCDPA) from [25]. © 2019 IEEE.

supply of 0.6 V and output power of –19 dBm, the DE is 10.7%.

It can be concluded that the PA power consumptions in ULP transmitters can be extremely low. For example, an efficiency of 40% at –20-dBm output power means that the PA only consumes 25 µW. Achieving a good overall TX efficiency at such low output powers is not a trivial task. The local oscillator, modulator and other TX circuitry need to consume ultra-low power. If such a 25-µW PA was combined with the 379-µW carrier generation mechanism used in [25], the TX efficiency would be at most 2.5%. On the other hand, if it was combined with the 22-µW LO of [11], an efficiency of 21% would be achieved. In addition to optimizing the power consumed in carrier generation, very specific modulation schemes need to be used to avoid further efficiency degradation due to the power consumed by the modulator circuitry.

2.5.4 State-of-the-art ULP transmitter implementations

Pandey et al. [11] presented the transmitter architecture consisting of a multi-phase LF-LO and an ECPA. The transmitter schematic is shown in Fig. 2.13(a). A stable reference frequency of 44.5-MHz is generated using a Pierce type XO.
Two cascaded injection-locked ROs (ILRO1 and ILRO2 in Fig. 2.8(b)) are utilized to generate the nine LO signal phases that are fed to the ECPA. Using edge-combining, the XO frequency is multiplied by a factor of nine as depicted earlier in Fig. 2.9. BFSK is performed by pulling the quartz reference clock using a switched capacitor, placed in parallel with the crystal, as shown in Fig. 2.13(a). The frequency is pulled by $\Delta f = 20$ kHz only as the frequency deviation is limited by the shunt capacitance of the crystal, parasitic capacitance and the oscillator startup requirements. However, as the ECPA multiplies the XO frequency by nine, the carrier frequency is shifted by $9 \cdot \Delta f = 180$ kHz. The transmitter consumes $24$ $\mu$W in the carrier generation and modulation including the XO. The output power is $20$ $\mu$W (i.e. $-17$ dBm) and the DE of the ECPA is $30\%$. The total power consumption is $90$ $\mu$W and the global efficiency (GE) is thus $20$ $\mu$W/$90$ $\mu$W $\approx 22\%$. The data rate is $200$ kbps and an EPB of $450$ pJ/bit is achieved. A spectrum of the modulated BFSK signal is not presented in the paper and, thus, the spectral quality of the modulated signal cannot be assessed. However, the first harmonic of the TX outputs when the bits 0 and 1 are transmitted is shown in Fig. 2.13(b). In an uplink measurement, the transmitter was able to deliver data over a distance of 5 meters to a Texas Instruments CC1101, a commercial off-the-shelf receiver, with a BER $< 10^{-3}$. 

Figure 2.13. (a) Transmitter architecture with injection-locked ring oscillators and an edge-combining PA, and (b) spectra of the output showing the two BFSK frequencies in [11]. © 2011 IEEE.

Figure 2.14. Spectrum with OOK-modulated signal in [23]. © 2018 IEEE.
Srivastava et al. [23] presented a transmitter consisting of an LF-LO and an ECPA but they used OOK instead of BFSK. Thus, the XO frequency pulling is omitted and the carrier is switched on and off by modulating the supply of the ECPA. The frequency of their Pierce type XO is 44.56 MHz and nine LO phases are generated using a single injection-locked RO. The transmission frequency is the XO frequency multiplied by nine, i.e. 401 MHz. The peak output power is $-17.6$ dBm but, with a modulated signal with a 50% duty cycle, the output power is reported to be $-24$ dBm. The data rate is 200 kbps and the transmitter consumes 71 $\mu$W from a 1-V supply. The GE and EPB are 5.6% and 360 pJ/bit, respectively. The output spectrum with OOK data at 200 kbps is shown in Fig. 2.14. The transmitter is able to deliver data wirelessly over a distance of 2 meters to a Texas Instruments CC1101 with a BER $< 10^{-3}$.

Tsai et al. [13] presented a 20-Mbps BPSK transmitter implemented using an LF-LO and an ECPA. To enable phase-shifting, instead of using a single-ended ILRO, they use a differential injection-locked low-frequency ring VCO, shown in Fig. 2.15(a). The nine output phases $P_1$–$P_9$ are equivalent to $A_1$–$A_9$ of Fig. 2.9. Each of the delay stage outputs is connected to a phase multiplexer. The TX data bit defines which phase of each delay stage output is directed to the ECPA. Looking at Fig. 2.9(a), it can be seen that inverting all the signals $A_1$–$A_9$ (i.e. $P_1$–$P_9$) also inverts the phase of the generated RF carrier. Therefore, a BPSK-modulated signal can be generated. The spectrum of the generated BPSK-modulated signal is shown in Fig. 2.15(b). The null-to-null bandwidth is 40 MHz due to the data rate of 20 Mbps. The supply voltages of the LO and ECPA are 0.8 V and 0.2 V, respectively. An external 44.45-MHz reference is used for injection-locking the ring VCO. The output power, TX power consumption and EPB are $-15$ dBm, 330 $\mu$W and 16.5 pJ/bit, respectively. The GE is 9.6%. The PA power consumption and drain efficiency are 147 $\mu$W and 21.5%, respectively.

Ma et al. [27] presented an OOK transmitter that exploits edge-combining and contains a class-C PA, shown in Fig. 2.16(a). A five-stage single-ended current-starved RO is injection-locked to the 5th harmonic of an external 16-MHz reference clock to produce an 80-MHz LO signal. An edge-combiner (EC) multiplies the LO frequency by five, producing a 400-MHz carrier, and drives the PA through a buffer. The supply voltage is 0.6 V, the TX consumes 160 $\mu$W and the output power is $-17$ dBm with OOK-modulated data. The data rate and
EPB are 1 Mbps and 160 pJ/bit, respectively. It can be calculated that the GE is 12.5%. The output spectrum is shown in Fig. 2.16(b). However, the span is only 750 kHz while the null-to-null bandwidth with OOK is expected to be twice the bit rate [6], i.e. 2 MHz. Therefore, the nulls and side lobes are not visible.

Paidimarri et al. [12] presented a transmitter capable of generating OOK, BPSK, minimum-shift keying (MSK) and Gaussian MSK (GMSK) signals. In contrast to many other sub-mW narrowband transmitter works, they use pulse shaping to suppress out-of-band PSD. The architecture is a modification of the direct-modulation architecture of Fig. 2.7(a). The carrier is generated using an FBAR oscillator. Three FBAR oscillators are included to enable use of three channels near 2.4 GHz. The oscillator outputs are multiplexed to the PA through a buffer circuit that prevents the PA from loading the oscillators. The buffer circuit contains two signal paths of which the other is delayed by 180°. This enables toggling of the signal phase which enables performing BPSK modulation. In the case of OOK, toggling of the phase is used to suppress the spectrum peak\(^8\) at the carrier frequency. It could otherwise cause harmful interference as, because of the pulse shaping, it would get mixed to other bands [12].

Paidimarri et al. utilize the push-pull PA of Fig. 2.12(a) that contains an on-chip matching network. The capacitors of the matching network can be switched to adjust the PA load impedance for controlling the output power. This enables pulse shaping which improves spectral efficiency [12]. The data rate is 1 Mbps. A 0.7-V supply is used with the RF circuits and a 1.0-V supply with the digital switches in the multiplexers and capacitor banks. Each FBAR oscillator consumes 150 µW and oscillator startup time is less than 4 µs. With OOK, BPSK and GMSK, the EPBs are 440, 530 and 550 pJ/bit, respectively, and output powers are –12.5, –11 and –10 dBm, respectively. From the EPBs and data rate, it can be calculated that the power consumptions are 440, 530 and 550 µW. Furthermore, from the power consumptions and output powers, it can be calculated that the GEs during the transmission of modulated data are 12.8%, 15.0% and 18.2%, respectively. The peak GE is 28.6% with the PA supply

\(^8\)This spectrum peak at the carrier frequency is not necessarily present if the LO is switched off during the transmission of zeros. Switching the LO on and off generally scrambles the phase which suppresses this peak.
of 0.7 V. However, the GE is lower with modulated data, assumingly because the PA output power is not at maximum all the time with the pulse shaping. The envelope shaping may decrease the efficiency but it improves the spectrum significantly. Fig. 2.17(a) shows an example of a generated BPSK signal with pulse shaping. The time-domain waveforms of the MSK or GMSK signals are not presented but the spectra are shown in Fig. 2.17(b). This also depicts how well the pulse shaping decreases the PSD of the side lobes of the spectrum.

Paidimarri et al. say [12] that the transmitter is able to generate MSK and GMSK signals. At this point, it can be noted that coherent MSK is able to achieve the error performance of binary PSK [5], [6]. However, to generate an MSK signal that reaches this performance, it is not enough to simply shift the frequency by $\pm 0.25 \cdot R_b$. Instead, simultaneous and precise control of both the transmit frequency and phase is required. Regarding their MSK implementation, Paidimarri et al. only discuss about shifting the frequency by $\pm 0.25 \cdot R_b$ and do not present any means for precisely controlling the phase at the same time nor do they discuss coherence. Therefore, there is reason to suspect that the error performance of coherently detected MSK [5], comparable to BPSK, cannot be achieved with the generated MSK signals. Due to uncertainties related to the error performance of the generated MSK signals, the performance in the MSK modes in [12] is omitted in any transmitter comparison in this thesis.

Mercier et al. [28] presented a PO-based OOK/FSK transmitter for the 2.4 GHz band. Fig. 2.18(a) shows the system schematic with the PO on the right. The power oscillator consists of a cross-coupled pair (transistors M1 and M2) connected to a LC tank consisting of a tunable on-chip capacitor and an off-chip center-tapped loop antenna. The oscillation amplitude can be controlled using the High-Vt switches $Mf[5:0]$. OOK is performed by toggling the amplitude. To perform BFSK, the modulator toggles the control word of the LC tank capacitor in accordance with the transmit data. MSK mode is presented as well but, as was the case with the transmitter of Paidimarri et al., the MSK mode been implemented only by shifting the carrier frequency by $\pm 0.25 \cdot R_b$ and no phase control or coherence is discussed. Presumably the generated MSK signal is not able to reach the error performance of coherently detected MSK. Because of the uncertainty with the error performance, the performance in the MSK mode is not considered in the transmitter comparisons in this thesis. In OOK and BFSK modes, the transmitter consumes 191 and 374 $\mu$W from a 0.8-V supply.
respectively, and the output powers are –29 and –26 dBm, respectively. With a data rate of 5 Mbps in each mode, the EPBs are 38 and 75 pJ/bit, respectively. From the output powers, EPBs and data rate, it can be calculated that the GE is 0.7% with both modulation schemes. The standby power consumption of the transmitter is ultra-low, only 39.7 pW.

2.5.5 Discussion

Table 2.1 summarizes the measured performance of the state-of-the-art sub-mW narrowband radio transmitters discussed in Section 2.5.4. The new energy efficiency FOM discussed in Section 2.4 and [I] has been calculated for each transmitter using (2.26). The table includes the parameters required for the FOM calculation. Furthermore, the table shows the link strengths that have been calculated using (2.28). More low-power direct-modulation transmitters can be found, for example, in references [10], [29], [30], [31], [32], [33] and PO-based transmitters in [10] and [20]. A more comprehensive energy efficiency comparison is provided in Section 5.3 after the narrowband transmitter implementations of this work and their measured performance have been discussed.

Among the discussed state-of-the-art transmitters, and among previously published ULP narrowband transmitters in general, the lowest power consumptions have been achieved by exploiting edge-combining in the carrier synthesis. This has resulted in power consumptions below 100 µW and also decent global efficiencies up to 22% with an output power of –17 dBm as is shown by Pandey et al. in [11]. Srivastava et al. and Ma et al. use an edge-combining architecture with OOK and they achieve lower EPBs than Pandey et al. but, due
Table 2.1. Comparison of prior ultra-low power narrowband transmitters

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<td>9.6%</td>
<td>12.5%</td>
<td>12.8% / 15.0%</td>
<td>0.7% / 0.7%</td>
</tr>
<tr>
<td>$P_{standby}$ (nW)</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>-</td>
<td>0.0397</td>
</tr>
<tr>
<td>Pulse Shaping</td>
<td>No</td>
<td>No</td>
<td>No</td>
<td>No</td>
<td>Yes</td>
<td>No</td>
</tr>
<tr>
<td>EPB (µJ/bit)</td>
<td>450</td>
<td>360</td>
<td>16.5</td>
<td>160</td>
<td>440 / 530</td>
<td>38 / 75</td>
</tr>
<tr>
<td>$P_n$ (dBm)</td>
<td>-120.8</td>
<td>-120.8</td>
<td>-100.8</td>
<td>-113.9</td>
<td>-113.9</td>
<td>-106.9</td>
</tr>
<tr>
<td>$\gamma_{avg,req}$ (dB)</td>
<td>13.4</td>
<td>13.1</td>
<td>9.6</td>
<td>13.1</td>
<td>13.1 / 9.6</td>
<td>13.1 / 13.4</td>
</tr>
<tr>
<td>Link strength (dB)</td>
<td>90.4</td>
<td>83.7</td>
<td>76.2</td>
<td>83.8</td>
<td>88.3 / 93.3</td>
<td>64.8 / 67.5</td>
</tr>
<tr>
<td>FOM (dBJ/bit)</td>
<td>-183.9</td>
<td>-178.2</td>
<td>-184.1</td>
<td>-181.7</td>
<td>-181.8 / -186.0</td>
<td>-168.9 / -168.7</td>
</tr>
</tbody>
</table>

1) During transmission of modulated data.
2) $P_n = k \cdot T \cdot BW_N$ where $T = 298$ K is noise temperature and $BW_N$ is noise bandwidth. With OOK, BPSK and BFSK, $BW_N$ is equal to symbol rate, i.e. data rate [17].
3) For BER = $10^{-5}$, 9.6, 13.1 and 13.4 dB with BPSK, OOK and BFSK, respectively.
4) Link strength and FOM calculated with (2.28) and (2.26), respectively.

To the lower power efficiencies, their FOMs are worse. Tsai et al. utilize a significantly high data rate of 20 Mbps which results in the lowest EPB among these works. Furthermore, they utilize BPSK and gain an advantage in the SNR requirement ($\gamma_{avg,req}$). Nevertheless, due to the high data rate, the bandwidth is wide, the noise power in the noise bandwidth is in the higher end, and the link strength is low. The energy efficiency improvement enabled by the use of BPSK is canceled by the lower power efficiency and, consequently, the energy efficiency FOM achieved by Tsai et al. is in the same range with that of Pandey et al. The transmitters in this comparison that exploit edge-combining have been implemented for carrier frequencies near 400 MHz and the efficiency is partly enabled by the low carrier frequency. Ring oscillators are used in these transmitters to generate the multiple phases fed to the edge-combiner circuitry. The carrier generation with a ring oscillator may require more power at higher carrier frequencies.

Mercier et al. presented a PO-based transmitter operating at 2.4 GHz. The transmitter features a relatively low power consumption, a particularly low standby power, and one of the lowest EPBs. However, the transmitter suffers from a low global efficiency of 0.7%. Consequently, the energy efficiency FOM is the highest among these works. Similar PO-based transmitters are presented in [20] and [III] and, in these works, the power efficiencies are low as well.
Moreover, the carrier frequency of a PO-based transmitter is susceptible to strong interferers and nearby physical objects [II]. These features may make the PO-based architecture less attractive for radio products that should be energy-efficient and comply with spectrum regulations. The PO-based architecture has, however, enabled very compact-sized systems. The transmitter chip of Mercier et al. with the on-PCB loop antenna has been demonstrated on a PCB with dimensions of 11-by-9 millimeters. Another miniature PO-based design is shown in [21] where a PO-based transmitter with a carrier frequency of 6.3 GHz has been fully integrated on a 1.4 mm-by-1.4 mm chip including the loop antenna.

The above works that have achieved low powers and EPBs utilize regular OOK, BPSK or BFSK without spectral shaping. Paidimarri et al. [12] implemented a transmitter capable of pulse shaping which improves the spectrum of the transmitted signal. The power consumption and EPB are in the higher end among these works. On the other hand, also the output power is high which, together with the conservative data rate and noise bandwidth, enables the highest link strength. The power efficiency is in the higher end and, because BPSK is used as well, the transmitter achieves the best energy efficiency FOM, i.e. the link strength is particularly high with respect to the EPB.

This was a brief comparison of state-of-the-art ULP transmitters and their energy efficiency FOMs. The new FOM expresses how high the link strength is with respect to the EPB. In this way, the FOM reflects the true energy-related challenge in transmitters – it is difficult to increase the link strength without increasing the EPB and, conversely, to reduce the EPB without reducing the link strength. Achieving a good FOM generally requires the use of power-efficient circuitry and, furthermore, the use of a modulation scheme that requires low SNR per bit to achieve the targeted error probability [I]. While low power consumption and EPB are generally desirable, other performance metrics should not be neglected in energy efficiency comparison. The new FOM accounts for multiple energy-related parameters. In this way, it provides a better overall view of the transmitter performance in terms of energy than the power consumption, power efficiency, EPB or the transmitter energy FOMs that have been used before. While this brief comparison only consisted of six state-of-the-art transmitters, a more extensive energy efficiency comparison with a greater amount of transmitters is provided in Section 5.3 after the sections that describe the narrowband transmitters implemented as a part of this work.

2.6 Summary

The choice of modulation scheme can have a significant impact on the power consumption of a transmitter. It defines the characteristics of the RF signal that a transmitter must generate. This, in turn, defines what kind of circuitry must be implemented for generating the signal. Due to differing SNR requirements of modulation schemes, the choice also affects the amount of output power that is
required to achieve the desirable link range.

Within prior published ULP narrowband transmitters, the lowest power consumptions have been achieved using OOK, BPSK and BFSK. These are modulation schemes that encode one bit per symbol that is transmitted in one baseband clock cycle the length of $T_{BB}$. The main benefit of OOK, BPSK and BFSK is that the RF signal can be generated without an excessive amount of high-power circuits such as linear PAs, linear mixers, phase-locked loops and digital-to-analog converters (DACs). The simplest OOK, BPSK and BFSK transmitter implementations only consist of a modulator and either a power oscillator or a combination of an LO and a PA. For this, some popular ULP carrier synthesis methods and power amplifier solutions were reviewed. In addition to enabling low transmitter complexity, OOK, BPSK and BFSK require a relatively low SNR per bit to achieve a given error probability. This enables lower output power and, hence, PA power consumption. Moreover, if OOK or BFSK is transmitted to a noncoherent receiver, carrier synchronization need not be performed. The drawback of these binary modulations is that they produce somewhat much out-of-band power unless pulse shaping is used. Moreover, their spectral efficiency is low and, for a high data rate, these binary modulations require a wider bandwidth compared to, for example, M-ASK, M-PSK or QAM.

Transmitter energy efficiency and prior transmitter energy FOMs were discussed, and it was concluded that prior transmitter publications have lacked a rational FOM. The new FOM, introduced in [I], was then reviewed. By referencing the link strength to the EPB, the new FOM provides a more comprehensive overall view of transmitter performance in terms of energy compared to other metrics such as the power consumption, power efficiency, EPB or the earlier transmitter FOMs. A few state-of-the-art ULP transmitters and their energy efficiency FOMs were compared. It was brought up that achieving good energy efficiency requires not only power-efficient circuitry but also the use of an energy-efficient modulation scheme. Furthermore, it was concluded in Section 2.3 that it may be more energy-efficient to transmit data using a high baseband clock frequency $f_{BB}$. However, with a high $f_{BB}$, more output power is required for a given SNR.

M-ary modulation using PPM and DPPM could be utilized for achieving improved transmitter energy efficiency. The next chapter discusses the properties, benefits and drawbacks of these two pulse-position modulation schemes. In Chapter 4, these schemes are compared with OOK, BPSK and BFSK in a new way to evaluate which modulation scheme could enable the highest energy efficiency, i.e. enable a given link strength with the least energy consumed per bit. The results suggest that PPM and DPPM could be more energy-efficient. This is also seen in the results achieved with the narrowband DPPM transmitters implemented as a part of this thesis that are discussed in Chapter 5.
3. M-ary PPM and DPPM

While ULP narrowband transmitters have generally used binary modulation schemes, it could also be beneficial to use M-ary modulation techniques [4]. This topic has been addressed in [I] and [II]. When using an M-ary modulation scheme, more bits are encoded per transmitted symbol. The TX front-end transmits an RF waveform in time $T_{BB}$ but now, instead of one bit, multiple bits are conveyed by that waveform. Thus, M-ary encoding decreases the active time per bit of the high-power blocks of the RF front-end. The potential benefit of M-ary modulation can be seen by writing an equation for the EPB. The transmitter front-end consumes power $P_{RF}$ over a time interval of one baseband clock cycle (i.e. $T_{BB}$) as a symbol is transmitted. The symbol conveys $B$ bits. Thus, as discussed in [II], the energy consumed per bit can be written as

$$EPB = \frac{P_{RF} \cdot T_{BB}}{B}.$$  \hspace{1cm} (3.1)

This equation suggests that the EPB reduces if $B$ is increased, i.e. more bits are encoded per symbol. However, the topic is somewhat complex because $P_{RF}$ is a factor that is affected by $B$ and the chosen modulation scheme. Firstly, $B$ affects it because increasing the output constellation size generally requires changes in the transmitter circuitry. Secondly, different modulation schemes generally require different SNRs to achieve a given error probability. This affects $P_{RF}$ through the PA power consumption – the greater the required SNR, the more output power is needed unless the noise bandwidth is altered.\(^1\) Nevertheless, the equation implies that a low EPB is achieved if multiple bits are encoded per symbol and the modulation scheme is such that 1) the RF signal with high $B$ (i.e. a large constellation) can be generated with low-power circuitry, and 2) it requires a low SNR per bit for the targeted error probability [II]. M-ary PPM and DPPM are considerable modulation schemes for energy-efficient ULP transmitters because they fulfill both requirements.

The advantages of PPM and DPPM have been discussed in [I] and [II]. With\(^1\)To meet a given SNR requirement, the achievable output signal SNR can also be altered through the baseband clock frequency $f_{BB}$ which impacts the noise bandwidth (see Sections 2.3 and 3.3.3). We consider a fixed $f_{BB}$ and noise bandwidth here. Therefore, changes in SNR or SNR requirements correspond to changes in signal power.
PPM, the data is encoded in the temporal position of a pulse within a symbol period. The benefit of PPM is that the signal resembles an OOK signal and can be generated in a similar fashion with a low-complexity low-power transmitter architecture such as the direct-modulation and the PO-based architectures that were discussed in Section 2.5. Noncoherent reception is possible which avoids carrier synchronization between the transmitter and the receiver. The modulation can be performed with relatively low power. The modulator can be, for instance, a low-power digital block based on a counter that operates at the baseband clock frequency $f_{BB}$ and defines the timing of the transmitted RF pulses. With PPM, the penalty of increasing $B$ is, however, that the signal “expands” in the time domain, assuming that $f_{BB}$ and $T_{BB}$ are maintained constant. Hence, the cost of high $B$ is decreased data rate. On the other hand, this decreases the average power consumption significantly. DPPM has several of the same benefits as PPM but allows a faster data rate than PPM with equal $B$ and $f_{BB}$.

This work focuses on PPM and DPPM. Other options for M-ary encoding are, for example, M-ary ASK, PSK, QAM and FSK. There are, however, some challenges related to all of these schemes. To maintain error performance, M-ary ASK, PSK and QAM generally require an increased SNR per bit as $B$ is increased [4], [6]. This implies increased PA power consumption and increased $P_{RF}$ when $B$ is increased. Conversely, orthogonal modulations such as M-ary FSK and PPM require less SNR per bit as $B$ is increased [4]. Moreover, generation of, for example, an M-ary PSK or QAM signal may require excessive high-power circuits [10] such as linear PAs or circuitry for generating multiple RF signal phases. This, again, may lead to increased TX circuit power consumption, i.e. increasing $B$ may require increased $P_{RF}$, making it more difficult to achieve good energy efficiency. The main benefits of PPM and DPPM over these modulation schemes are the low SNR per bit required for a given error probability and the possibility to use very simple circuitry. Moreover, as opposed to M-FSK that operates on multiple tones, PPM and DPPM operate on a single carrier tone which simplifies frequency calibration. Furthermore, with FSK, the signal bandwidth increases as $B$ is increased unless $f_{BB}$ is also decreased.

### 3.1 Pulse-position modulation

PPM is an orthogonal modulation [34]. Compared to the conventional binary modulation schemes utilized in ULP transmitters, use of PPM offers energy efficiency improvement because of 1) decreased active time per bit of the RF front-end, and 2) a relaxed SNR-per-bit requirement. The former follows from the fact that one symbol carries multiple bits instead of one. The latter is a consequence of signal orthogonality which enables improved error performance.
with respect to the SNR per bit. It is worth noting that M-ary PPM requires a higher peak output power compared, for instance, to BPSK and BFSK. However, if sufficiently many bits are encoded per symbol with PPM, the required amount of transmitted energy per bit is lower compared to these modulation schemes.

With M-ary PPM, $B$ bits are encoded per symbol. The symbol duration consists of $M = 2^B$ slots the length of $T_{BB}$. A pulse is placed to one of these slots to define which $B$-bit value the symbol represents. The slot with the pulse is called an “on” slot and the slots without the pulse are called “off” slots. Fig. 3.1 presents the baseband signal waveforms of 4-PPM. With 4-PPM, the value of $M$ is 4 and $B = 2$ bits are encoded per symbol. Thus, the length of the symbol is four baseband clock cycles (i.e. $4T_{BB}$). Only one pulse is transmitted per two bits and the active time of the front-end is $\frac{1}{2} \cdot T_{BB}$ per bit. This is equal to OOK. With BPSK, on the other hand, the active time per bit is $T_{BB}$. In general, the active time per bit with M-PPM is $\frac{1}{B} \cdot T_{BB}$, the active ratio is $R_{t,PPM} = 1/B$ and the active time per bit decreases inversely proportionally to $B$ [I]. If the LO of the transmitter can be duty-cycled efficiently, this enables significant reduction of LO energy consumption per bit. For example, a free-running VCO or DCO could be used as the LO with negligible energy overhead in duty-cycled mode. If frequency accuracy is a concern, the frequency could be calibrated before transmission by utilizing a PLL in a similar fashion to [21].

As the duration of a slot is $T_{BB}$, the duration of a symbol with M-ary PPM is

$$T_s = 2^B \cdot T_{BB}. \quad (3.2)$$

With $B$ bits encoded per symbol and transmitted per $T_s$, the data rate is

$$R_b = \frac{B}{T_s} = \frac{B}{2^B \cdot T_{BB}} = \frac{B \cdot f_{BB}}{2^B}. \quad (3.3)$$

Thus, the data rate is lower than with the conventional binary modulation schemes (i.e. OOK, BPSK and BFSK) whose data rate is $f_{BB}$. In other words, the energy efficiency improvement comes at the cost of data rate. The transmitter transmits the carrier for $T_{BB}$ once per $T_s$. Hereby, the duty cycle is

$$R_{DC} = \frac{T_{BB}}{T_s} = \frac{1}{2^B}. \quad (3.4)$$

This enables a low average power consumption.
3.1.1 Error performance of PPM

Orthogonal signaling possesses a favorable property in terms of transmitter energy efficiency – the more bits are encoded per symbol, the less SNR is required per bit to achieve a given BER. According to [4], with orthogonal modulation schemes, the theoretical probability of a symbol error in an AWGN channel with noncoherent detection is

$$SER_{\text{orthogonal}} = \sum_{n=1}^{M-1} \frac{(-1)^{n+1}}{n+1} \binom{M-1}{n} \cdot \exp\left(-\frac{n \log_2 M E_b}{n+1 N_0}\right). \quad (3.5)$$

The theoretical BER can be derived by multiplying this by $(2^{B-1})/(2^B - 1)$ [4] which yields

$$BER_{\text{orthogonal}} = \frac{2^{B-1}}{2^B - 1} \sum_{n=1}^{M-1} \frac{(-1)^{n+1}}{n+1} \binom{M-1}{n} \cdot \exp\left(-\frac{n \log_2 M E_b}{n+1 N_0}\right). \quad (3.6)$$

Equations (3.5) and (3.6) apply not only to PPM but orthogonal modulation schemes in general. Fig. 3.2 and Fig. 3.3 show the theoretical symbol and bit error ratios, respectively, of noncoherently detected PPM with $B$ values from

![Figure 3.2. SER of noncoherent PPM.](image)

![Figure 3.3. BERs of noncoherent OOK, coherent BPSK and noncoherent PPM.](image)
M-ary PPM and DPPM

1 to 6, obtained using (3.5) and (3.6). The more bits are encoded per symbol (i.e. per transmitted RF pulse), the less SNR is required per bit. For reference, Fig. 3.3 also shows the BERs with noncoherently detected OOK and coherently detected BPSK, obtained with (2.4) and (2.5). With values of \( B \) greater or equal to 5, PPM outperforms coherent BPSK at BER values below \( 10^{-2} \). The error performance is equal or better compared to OOK and BFSK with \( B \geq 2 \). To achieve a BER of \( 10^{-5} \) in an AWGN channel, OOK, BPSK and PPM with \( B = 6 \) require 13.1, 9.6 and 6.7 dB of SNR per bit, respectively. No separate plot is drawn for noncoherent BFSK as it is an orthogonal modulation scheme as well\(^3\) and its BER is equal with 2-PPM (\( B = 1 \)). Hence, BFSK requires 13.4 dB for \( \text{BER} = 10^{-5} \).

Two conclusions can now be made about PPM. According to (3.1), M-ary PPM can reduce the energy consumed per bit by the transmitter – by increasing \( B \), the active time per bit reduces and, hence, the EPB reduces. M-ary PPM is also efficient in terms of SNR: according to Fig. 3.3, the greater the value of \( B \), the less SNR is required per bit to achieve a given bit error probability. This translates to conserved energy due to reduced PA energy consumption per bit. The combined effect of the decreased active time per bit and lower required SNR per bit is analyzed in Chapter 4. For the analysis, it is beneficial to express the SER as a function of the SNR \( \gamma \) as opposed to the SNR per bit. As each symbol carries \( B \) bits, the relationship between the SNR per bit and SNR is \( E_b/N_0 = \gamma/B \). Substituting this to (3.5) yields

\[
\text{SER}_{\text{orthogonal}} = \sum_{n=1}^{M-1} \frac{(-1)^{n+1}}{n+1} \binom{M-1}{n} \cdot \exp\left(-\frac{n \cdot \gamma}{n+1}\right). \tag{3.7}
\]

### 3.2 Differential pulse-position modulation

Differential PPM [8], [9], [I], [II], [IV] is a modulation similar to PPM but it offers a greater data rate and avoids symbol synchronization. A DPPM symbol is obtained from a PPM symbol by removing the “off” slots that follow the “on” slot [9]. Fig. 3.4 shows 4-DPPM symbols as depicted in [I]. The data content of a DPPM symbol is defined by the position of the “on” slot within the symbol period. When the symbols are transmitted sequentially, the data is practically encoded

\(^3\)Assuming properly [4], [7] chosen transmit tones.
M-ary PPM and DPPM

also in the time delays between the pulses. Therefore, the demodulation can be performed by calculating the delays between the pulses (i.e. between the “on” slots) as discussed, for instance, in [II] and [IV].

As the symbols are shorter with DPPM, the average data rate with DPPM is greater compared to PPM with equal $B$ and $f_{BB}$. When a DPPM symbol represents an integer data word with value $W$ in the range $[0, 2^B-1]$, the duration of the symbol is $(W+1)\cdot T_{BB}$. The duration of the shortest and longest symbol are $T_{s,\text{min}} = T_{BB}$ and $T_{s,\text{max}} = 2^B \cdot T_{BB}$, respectively. As per [IV], the average symbol time with equiprobable symbols is

$$T_s = \frac{2^B + 1}{2} \cdot T_{BB}. \tag{3.8}$$

With $B$ bits encoded per symbol and transmitted per $T_s$, the average data rate with DPPM is [IV]

$$R_b = \frac{B}{T_s} = \frac{2^B}{2^B + 1} \cdot f_{BB}. \tag{3.9}$$

Moreover, an interesting figure regarding ULP transmitters is the average duty cycle. One pulse the length of $T_{BB}$ is transmitted per $T_s$ on average. Hence, the average duty cycle is

$$R_{DC} = \frac{T_{BB}}{T_s} = \frac{2}{2^B + 1}. \tag{3.10}$$

The average duty cycle decreases rapidly as $B$ increases. Therefore, DPPM offers low average power consumption.

DPPM reception does not require symbol synchronization that is required for PPM detection [9]. This follows from the fact that the data is effectively encoded in the delays between the pulses. This allows a receiver to demodulate the data by calculating the times between the pulses. However, looking at Fig. 3.4, it is obvious that, if the transmit data contains several consecutive symbols representing the smallest data word (002 in the case of 4-DPPM), there are no delays between the pulses at all. Now, a problem arises if the transmitted pulses are clocked with a ring oscillator or other oscillator with uncertainty in the frequency. With a long stream of inseparable pulses, it may be impossible for a receiver to know how many pulses the stream contains. Thus, to make the pulses separable, a guard slot may be included in the symbols. DPPM with guard slots is discussed in the next section.

### 3.2.1 DPPM with guard slots

A guard slot is an additional “off” slot added to the beginning of each symbol. 4-DPPM symbols with guard slots are depicted in Fig. 3.5. With the guard slots, two pulses are always separated by a gap. This helps a receiver to distinguish the pulses from each other. However, this also increases the symbol time and decreases the data rate. As each symbol is now $T_{BB}$ longer, the duration of the
shortest and longest symbol are $2^B \cdot T_{BB}$ and $(2^B + 1) \cdot T_{BB}$ [II]. The average symbol time with equiprobable symbols is [II]

$$T_s = \frac{2^B + 3}{2} \cdot T_{BB}. \quad (3.11)$$

Each symbol conveys $B$ bits and the average data rate is [II]

$$\bar{R}_b = \frac{B}{T_s} = \frac{2 \cdot B}{2^B + 3} \cdot f_{BB}. \quad (3.12)$$

The transmitter front-end is ideally active for $T_{BB}$ per $T_s$. The average duty cycle is thus

$$\bar{R}_{DC} = \frac{2}{2^B + 3}. \quad (3.13)$$

The DPPM modulators implemented in this thesis were intended to be clocked using ring oscillators. Due to the inaccuracy of this kind of a baseband clock generator, guard slots have been utilized. Hence, the following discussions about DPPM consider specifically DPPM with guard slots.

### 3.2.2 Differences between PPM and DPPM

DPPM is different from PPM in terms of error performance because PPM is orthogonal while DPPM is not. The orthogonality of PPM is related to the fact that each symbol consists of a fixed number of slots and the carrier is transmitted in a single slot only. DPPM is not orthogonal because the number of slots in a symbol is not constant. Hereby, the symbol and bit error probability of equations (3.5) and (3.6) that apply to orthogonal signals do not apply to DPPM.

To aid in understanding this difference, Fig. 3.6 depicts the principle of PPM reception and detection in the case of 4-PPM. There are $2^B = 4$ time slots and the signal energy is placed in one slot, in this case the third slot, to represent binary value $10_2$. Thus, the slot array is 0100. The PPM signal is received...
noncoherently with an envelope detector and the envelope detector output is ideally sampled once per slot. The envelope detector output is affected by noise for which the output consists of nonbinary values 1.01, 1.17, 0.66 and 0.84. The receiver interprets the slot with the highest output amplitude as the “on” slot. In this case, the highest envelope detector output amplitude is 1.17. This highest amplitude occurs correctly in the third slot and the receiver interprets this slot as the “on” slot.

A symbol error occurs at the receiver if, due to noise, the output amplitude in one of the “off” slots exceeds the amplitude of the “on” slot. In that case, an “off” slot is erroneously interpreted as the “on” slot. The probability of error here depends on the amount of slots – the more there are slots, the greater is the probability that the amplitude in one of the “off” slots is greater than the amplitude in the intended “on” slot. With DPPM, the amount of slots in a symbol depends on the data content and, obviously, the probability of error is not the same as with PPM. For this, the SER and BER equations of PPM as given in (3.5) and (3.6) cannot be applied to DPPM.

The error performance of DPPM is analyzed in Section 3.2.4. In particular, the error performance of packet-mode transmission is analyzed. This is due to problems related to continuous-mode data streaming with varying symbol durations. Continuous-mode DPPM reception is prone to insertion and deletion errors [35] – symbols may be added to or removed from a data stream if an “off” slot is interpreted as an “on” slot or vice versa. An insertion error occurs if an “off” slot is erroneously interpreted as an “on” slot. The result is that a symbol is inserted into the received symbol array. The effect of this on a received symbol stream is depicted in Fig. 3.7(a). As the data of the following symbol is interpreted based on the delay between this false “on” slot and the next “on” slot, this following symbol is erroneous. A deletion error occurs if an “on” slot is misinterpreted as an “off” slot in which case a symbol is deleted from the received symbol array. The effect on a received symbol stream is depicted in Fig. 3.7(b). Here, too, the following symbol is misinterpreted due to the change of delay between the pulses. These insertion and deletion errors shift all the subsequent symbols.

To reduce the problems associated with continuous-mode transmission, packet-mode transmission can be used instead. While insertion and deletion errors may still occur, they are isolated to a single packet and will not affect the subsequent packet. If continuous-mode transmission is necessary or the insertion and
deletion errors are a concern, PPM may be preferable due to the constant symbol time.

### 3.2.3 Packet-mode DPPM

In the intended mode of operation, the transmitters implemented in this thesis transmit data in packets. This isolates the effect of insertion and deletion errors to a single packet. Fig. 3.8 shows a simplified depiction of the packet format used in this work in [I] and [II]. For simplicity, an 8-DPPM data packet is shown with 12 bits encoded in four symbols. However, in the measurements, the transmitters of this work were mostly configured to transmit 48-bit 64-DPPM packets consisting of eight 6-bit symbols.

In Fig. 3.8, the packet begins with a start pulse. The timing of the pulse in the first symbol is referenced to it. $N_s$ symbols are transmitted per packet. Four symbols are transmitted and the start pulse is thus followed by four pulses. Guard slots are utilized and therefore there are $W+1$ “off” slots before each “on” slot where $W$ is the integer data word which is conveyed by the symbol.

Regarding the error probability with DPPM, one critical number is $N_O$, the number of “on” slots in a packet. With the start pulse, it is given by [II]

$$N_O = N_s + 1. \quad (3.14)$$

Another critical number is $N_Z$, the maximum number of “off” slots in a packet. It can be derived from $N_O$ and the maximum number of slots in a packet, $L_{p,\text{max}}$. With the start pulse and $N_s$ symbols, the maximum amount of slots per packet is [II]

$$L_{p,\text{max}} = 1 + N_s \cdot L_{s,\text{max}}, \quad (3.15)$$

where $L_{s,\text{max}}$ is the maximum number of slots in a symbol, given by [II]

$$L_{s,\text{max}} = 2^B + 1. \quad (3.16)$$

Substituting (3.16) to (3.15) yields [II]

$$L_{p,\text{max}} = 1 + N_s \cdot (2^B + 1). \quad (3.17)$$

According to (3.17), increasing the number of symbols increases the length of the packet linearly while increasing the word length (i.e. $B$) causes roughly an
exponential increase. The maximum number of “off” slots is given by [II]

\[ N_Z = L_{p,\text{max}} - N_O. \]  

(3.18)

By substituting (3.17) and (3.14), (3.18) becomes [I]

\[ N_Z = N_s \cdot 2^B. \]  

(3.19)

The equations for \( N_O \) and \( N_Z \) are useful in the error performance analysis of packet-mode DPPM, provided in the next section. With the transmitters implemented in this thesis, 48-bit packets are transmitted with \( N_s = 8 \) and \( B = 6 \). With this configuration, the maximum packet length given by (3.17) is \( L_{p,\text{max}} = 521 \) slots. The active ratio with DPPM is obtained with

\[ R_{t,\text{DPPM}} = \frac{N_s + 1}{N_s \cdot B}, \]  

(3.20)

because \( N_s + 1 \) pulses are required for transmitting \( N_s \cdot B \) bits with the packet-mode transmission considered in this work [I].

### 3.2.4 Error performance of packet-mode DPPM

A DPPM signal resembles an OOK signal – with both modulation schemes, there are time slots the length of \( T_{BB} \) where the carrier is either present or not present. The conventional way of decoding an OOK signal is to use a hard-decision approach: a decision threshold level is chosen and, if the envelope detector output signal is above the threshold, the bit is interpreted as 1, and otherwise as 0. This approach could be used with DPPM as well to decide whether a slot is an “on” or an “off” slot. However, attempting to set the decision threshold to the optimum level is even more challenging in the case of DPPM. In Section 2.2, it was discussed\footnote{Recall that the threshold is approximated as \( b_0 = \sqrt{2 + \gamma/2} \) [7] or \( b'_0 = \sqrt{2 + E_b/N_0} \).} that the optimal decision threshold with OOK depends on the signal-to-noise ratio \( \gamma \). Thus, for optimum error performance, an OOK receiver must estimate the SNR and adjust the threshold accordingly. With DPPM, the ratio of the “on” and “off” slots can be very low. Thus, estimating the SNR based on the received DPPM signal may be quite cumbersome and particularly at low SNR levels when the “on” slots are nearly inseparable from the “off” slots. To overcome this, a DPPM transmitter could transmit a pilot signal before the data content which a receiver could use for estimating the SNR. However, energy is required for transmission of the pilot and, in ultra-low power applications, it may be preferable to avoid its use. Instead, if the receiver utilizes soft-decision decoding [14], [35], [36], [III] such SNR estimation is avoided.

The operation of soft-decision decoding is similar to the detection of an orthogonal PPM signal which was depicted in Fig. 3.6. A noncoherent receiver with an envelope detector [7] can be used in the process. The receiver ideally samples the envelope detector (ED) output once per slot. With DPPM, symbol-by-symbol
detection is difficult because the symbol time varies according to data and a receiver does not know when a symbol starts and ends. Hence, we shall consider a packet-level soft-decision decoding (PL-SDD) scheme instead. The PL-SDD scheme has been discussed in [I] and [II]. Similar type of soft-decision decoding has been considered, for example, on symbol-level in [14]. In symbol-level decoding of, for instance, PPM, the receiver knows that one pulse is contained within one symbol consisting of $2^B$ slots. With DPPM, the receiver does not know the symbol durations and the same principle cannot be applied directly. Instead, this idea can be extended to packet-level: the receiver knows that there is one start pulse and one pulse for each symbol within one packet consisting of at most $L_{P,\text{max}}$ slots.

The operating principle of a PL-SDD receiver is depicted in Fig. 3.9. A transmitted DPPM baseband signal is shown in Fig. 3.9(b) and the resulting ED output in Fig. 3.9(a). In this simplified example, a packet consists of only $N_s = 2$ symbols with $B = 3$ bits encoded per symbol. The DPPM packet thus consists of three pulses: one start pulse and one pulse for each of the two symbols. The symbols are 100$_2$ and 111$_2$ and guard slots are utilized. The receiver samples the ED output amplitude once per slot during the maximum duration of a packet which in this case is $L_{P,\text{max}} = 1 + N_s \cdot (2^B + 1) = 19$ slots as given by (3.17). The receiver expects $N_O = N_s + 1 = 3$ “on” slots in the ED output. Hence, it interprets the slots with the $N_O = 3$ highest amplitudes as the “on” slots and the other ones as “off” slots. The time differences between these pulses are $6 \cdot T_{BB}$ and $9 \cdot T_{BB}$ and they are interpreted correctly as 100$_2$ and 111$_2$. As explained in [II], effectively, the decision threshold is set adaptively to the $N_O^{th}$ highest amplitude among all the slots, in this case the third highest peak.

For successful reception, it is enough if the ED output amplitudes in the $N_O$
"on" slots exceed the highest amplitude among the "off" slots. In Fig. 3.9(a), the ED output peak produced by the second DPPM pulse barely exceeds the exceptionally high peak in the tenth slot that is caused by noise. This noise peak significantly exceeds the mid-level of the "on" slots and "off" slots. With a hard-decision approach, the decision threshold would be much lower and the tenth slot would be falsely interpreted as an "on" slot. At moderate and high SNR levels, it is statistically probable that the lowest amplitude in the "on" slots is higher than the highest amplitude in the "off" slots. By effectively adapting the threshold according to the actual highest received signal peaks, a lower PER is achieved compared to the use of a statistically optimum threshold. The benefit of this PL-SDD approach is thus that the amplitudes in the "off" slots may exceed the mid-level of the amplitude range near which the threshold would be set in the hard-decision approach. Also the peaks produced by the DPPM pulses may occasionally fall below this mid-level without producing errors if the noise peaks happen to be low enough.

The error probability with the PL-SDD scheme in the AWGN channel can be calculated based on the probability distributions of an ED output as discussed in [II]. When a DPPM "on" slot is received, or bit 1 in the case of OOK, the ED output amplitude is Rician-distributed. As per [7], the ED output distribution (the probability that the output amplitude is $r$) is,

$$
p_1(r) = \frac{r}{\sigma^2}e^{-\frac{(r^2+A^2)}{2\sigma^2}}I_0\left(\frac{rA}{\sigma^2}\right), \quad (3.21)
$$

where $A$ is the signal envelope, $\sigma$ the standard deviation of additive white noise and $I_0(z)$ the modified Bessel function of the first kind and zeroth order [7], [14]. When a DPPM "off" slot is received, or bit 0 in the case of OOK, the output voltage is Rayleigh-distributed and the distribution [7] is given by

$$
p_0(r) = \frac{r}{\sigma^2}e^{-\frac{r^2}{2\sigma^2}}. \quad (3.22)
$$

Examples of Rayleigh and Rician distributions were shown in Fig. 2.4.

With the probability distributions (3.21) and (3.22), an equation can be derived for the packet error probability with the PL-SDD scheme. As per [II], it is

$$
P_{ERdppm} = \int^\infty_0 \frac{N_Z r e^{-r^2/2\sigma^2}}{\sigma^2} \left[ \int^r_0 \frac{r e^{-r^2/2\sigma^2}}{\sigma^2} dr \right]^{N_Z-1} \left\{ 1 - \left[ 1 - \int^r_0 \frac{r e^{-r^2/2\sigma^2}}{\sigma^2} I_0\left(\frac{rA}{\sigma^2}\right) dr \right]^{N_O} \right\} dr. \quad (3.23)
$$

Here, $N_O$ and $N_Z$ are the numbers of the "on" and "off" slots in a packet as given by (3.14) and (3.19). The derivation of (3.23) is shown in Appendix A. Regarding noncoherent reception using an envelope detector, (3.23) gives the probability that the envelope detector output amplitudes in the $N_O$ intended "on" slots exceed the amplitudes in the rest of the slots. Hence, it is the packet error probability with PL-SDD reception of DPPM. As a sidenote, computation
Figure 3.10. Packet error ratio with noncoherent OOK with hard-decision decoding and DPPM with the proposed packet-level soft-decision decoding scheme. \( \gamma \) is the SNR while carrier is being received (i.e. not the average SNR per bit). Obtained from [II], used under CC BY 4.0.

suggests that it also yields the symbol error probability with PPM if the numbers of “on” and “off” slots are set as \( N_O = 1 \) and \( N_Z = 2^B - 1 \), respectively.

For [II], (3.23) was evaluated using numerical integration. Fig. 3.10 shows the obtained packet error ratios with DPPM with the PL-SDD scheme with \( N_b = 48 \) bits per packet and \( B \) values of 2, 6, 8 and 12. The corresponding values of \( N_O \), obtained with (3.14), are 25, 9, 7 and 5, respectively. The values of \( N_Z \), obtained with (3.19), are 96, 512, 1536 and 16384. The results are plotted against \( \gamma = A^2/2\sigma^2 \) (i.e. the SNR while carrier is being received).\(^5\) For reference, the PER with noncoherent OOK with 48-bit packets is plotted. With the BER of noncoherent OOK given in (2.3), the PER with OOK has been calculated as

\[
PER = 1 - (1 - BER)^{N_b}.
\]  
(3.24)

According to Fig. 3.10, the DPPM schemes with up to \( B = 8 \) bits encoded per symbol achieve a lower PER than OOK when \( \gamma \) is greater than 11–12.5 dB. With 12 bits encoded per symbol, DPPM outperforms OOK when \( \gamma \geq 14 \) dB. The DPPM schemes with \( B \leq 8 \) require lower \( \gamma \) than OOK for a PER below \( 10^{-1} \). It is worth noting that \( \gamma \) is the SNR while carrier is received, not the average SNR per bit. Hence, the figure tells that noncoherent DPPM offers a similar error performance with noncoherent OOK with slightly less peak output power assuming equal noise bandwidths with the OOK and DPPM schemes.\(^5\)

Thus, without sacrificing packet error performance, an OOK transmitter can be converted to a DPPM transmitter just by replacing the OOK modulator with a DPPM modulator, clocked with the same \( f_{BB} \).

One important observation can be made about DPPM in Fig. 3.10: at low values of \( B \) such as the ones that have been plotted, the peak output power required\(^5\) (3.23) is a function of \( A \) and \( \sigma \), not a function of \( \gamma \). In evaluation of (3.23), one may choose \( A = 1 \) and then, as \( \gamma = A^2/2\sigma^2 \), \( \sigma \) is obtained using \( \sigma = \sqrt{A^2/2\gamma} = \sqrt{1/2\gamma} \).

\(^6\)It can be noted again that the SNR depends on both signal power and noise bandwidth. A higher SNR requirement does not directly mean that a greater output power is required – the SNR could also be improved through baseband clock frequency \( f_{BB} \) which impacts the noise bandwidth (see Sections 2.3 and 3.3.3).
M-ary PPM and DPPM

to achieve a certain PER does not increase strongly with $B$. In other words, a negligible PA power increment is required to maintain error performance as $B$ is increased. Looking at (3.1), this feature is advantageous – $B$ may be increased for a lower EPB and only a minor increment of PA power and $P_{RF}$ is required to avoid the consequent loss in error performance [II]. It can be seen in (3.12), however, that this also reduces the data rate. The transmitters designed in this work use $B$ values up to six to avoid excessively long packets.

Fig. 3.10 is informative regarding the peak output power. Regarding TX energy efficiency and PA energy consumption per bit, plotting the error performance against the average SNR per bit is more informative. The PERs of OOK, BPSK, BFSK and DPPM are plotted against this and compared in the next section. As a final point before that, a word could be said about the RX memory requirements. (3.17) suggests that plenty of memory is required to store all the envelope detector output amplitudes if $B$ is high and plenty of symbols are encoded per packet. Some sort of optimization of the PL-SDD reception algorithm could, however, perhaps be done. The RX is essentially interested only in the highest amplitude peaks in the ED output and their timing. This could possibly allow some memory reduction. This optimization is, however, out of the scope of this thesis and no optimized on-chip nor FPGA PL-SDD implementations were considered in this work. A software-radio-based PL-SDD DPPM RX was, however, implemented for testing the uplink with the DPPM transmitter presented in [II]. The details of this receiver are discussed in Section 5.1.4.

3.2.5 Packet error performance comparison with OOK, BPSK and BFSK

The PERs with OOK, BPSK and BFSK against the average SNR per bit are obtained by substituting their BER equations (2.4), (2.5) and (2.6), respectively, to (3.24). The PERs with $N_b = 48$ bits encoded per packet are plotted in Fig. 3.11. The plots with DPPM were obtained from those of Fig. 3.10 by scaling $\gamma$ to obtain the average SNR per bit. With DPPM, one RF pulse is transmitted per $B$ bits. In packet-mode, an additional start pulse is transmitted. The average SNR per bit is thus obtained from $\gamma$ by multiplying it by $(N_s + 1)/(N_s \cdot B)$. As the curves of Fig. 3.10 are plotted against the $\gamma$ in decibels, the DPPM plots against the average SNR per bit are obtained from the DPPM curves of Fig. 3.10 by adding $10 \cdot \log_{10}((N_s + 1)/(N_s \cdot B))$ to the horizontal axis values. With $B$ values of 2, 6, 8 and 12, the number of symbols per 48-bit packet, i.e. $N_s$, is 24, 8, 6 and 4, respectively. The corresponding shifts on the horizontal axis are approximately $-2.8$, $-7.3$, $-8.4$ and $-9.8$ dB, respectively. Additional plots with DPPM with $B = 1$ and $B = 4$ have been included in Fig. 3.11.

Fig. 3.11 shows the PERs with coherent BPSK and noncoherent OOK, BFSK and DPPM. The PER of DPPM here assumes use of PL-SDD at the receiver side. DPPM is an advantageous modulation in terms of output power and PA power consumption: for a PER below 0.01, DPPM with $B \geq 4$ requires less SNR
M-ary PPM and DPPM

0 2 4 6 8 10 12 14 16

10 -5
10 -4
10 -3
10 -2
10 -1
10 0

Packet size: 48 bits/packet

OOK
BPSK
BFSK
DPPM (B=1)
DPPM (B=2)
DPPM (B=4)
DPPM (B=6)
DPPM (B=8)
DPPM (B=12)

Figure 3.11. Theoretical packet error ratios with OOK, BPSK, BFSK and DPPM against the average SNR per bit. The PER of DPPM here assumes the use of PL-SDD in the receiver.

per bit than OOK, BPSK and BFSK, i.e. the conventionally used binary ULP modulations. In the likes of PPM, the required SNR per bit and active ratio (see (3.20)) both reduce as \( B \) is increased which makes M-ary DPPM schemes favorable options for ULP transmitters. While the transmitters implemented in this thesis mainly use \( B = 6 \), the figure shows that the required SNR per bit with \( B = 4 \) is still comparable to that of BPSK. In Chapter 4, it is shown that the use of both \( B = 4 \) and \( B = 6 \) could offer energy efficiency improvement compared to OOK, BPSK and BFSK.

3.3 Waveform-level PER simulations

This section discusses the waveform-level PER simulations of [I]. Related to the energy efficiency comparison of modulations discussed in [I] and Chapter 4, the goal of the simulations was to verify the validity of the utilized error probability equations and that the noise bandwidth is \( f_{BB} \) with all the compared modulation schemes. Note that by performing these kinds of simulations one may achieve a more profound understanding of how the choice of modulation scheme impacts the uplink performance and the energy efficiency of a transmitter. Sections 3.3.1 and 3.3.2 supplement the simulation description of [I] with additional simulation details. Section 3.3.3 provides additional discussion on the signal-energy-dependency of the error probability.

In [I], Mathworks MATLAB was used to simulate the PERs of BPSK, OOK, BFSK, 16-PPM, 64-PPM, 16-DPPM and 64-DPPM on waveform-level. Modulated RF signal waveforms representing 48-bit packets were generated and AWGN was added. The time-domain RF signals with the noise were fed to the receiver models which are described in Section 3.3.1. The signal sample rate, carrier frequency and baseband clock frequency were set to \( f_s = 10 \) GHz, \( f_c = 434 \) MHz and \( f_{BB} = 10 \) MHz, respectively. Thus, one baseband clock period was \( 1/f_{BB} = 100 \) ns, i.e. 1000 samples. With BFSK, the carrier tones representing bits 0 and 1 were \( f_c = 434 \) MHz and \( f_{c2} = 444 \) MHz, respectively. In the case of
OOK, PPM, DPPM and BFSK, noncoherent transmit signals were generated. It was assumed that the carrier phase becomes scrambled when, with OOK, PPM and DPPM, the LO is switched on and off, and, with BFSK, when the transmit tone is changed from $f_c$ to $f_{c2}$ or vice versa. Thus, the RF signal phase was randomized accordingly each time the amplitude or frequency was toggled.

### 3.3.1 Simulated receivers

Fig. 3.12 shows the receivers that were used in the PER simulations of [I]. The receivers are adaptations of the receivers and demodulators provided, for instance, in [5], [6] and [7]. Fig. 3.12(a) shows the simulated envelope-detector-based noncoherent OOK, PPM and DPPM receiver. The BPF is a filter matched [5] to the transmit frequency $f_c$. Thus, the finite impulse response (FIR) filter coefficients are a sine wave with frequency $f_c$, flat envelope and length $T_{BB}$, i.e. 1000 samples. The BPF is followed by an envelope detector which is formed by the full-wave rectifier and the integrator. The integration time interval is approximately equivalent to one RF sine cycle, $f_s/f_c \approx 23$ samples. The sampler samples the ED output with a period of $T_{BB}$.

The sampled ED output in Fig. 3.12(a) is fed to a modulation-scheme-dependent decision device. In the case of OOK, it is fed to a comparator. With OOK, the optimum threshold for the comparator is $b_0 = \sqrt{2 + \gamma^2}$ (see OOK BER discussion in Section 2.2) which depends on $\gamma = A^2/2o^2$. To optimize the threshold, a receiver must estimate $\gamma$. In the simulations, this estimation was done at each simulated...
SNR level by estimating the values of $A$ and $\sigma$ from the observed ED output distributions. The `mle()` function of MATLAB was used in the estimation. It returns the estimated parameters of a data set whose values conform to a known distribution. The sampled ED output is Rayleigh- and Rician-distributed (see Fig. 2.4) when zeros and ones are received, respectively. $\sigma$ can be estimated from both Rayleigh- and Rician-distributed data sets, and $A$ from a Rician-distributed data set. Thus, to estimate the optimum threshold, the reception of a decent amount of bits was simulated, $A$ and $\sigma$ were estimated from the ED output distributions using the `mle()` function, $\gamma = A^2/2\sigma^2$ was calculated based on the $A$ and $\sigma$ estimates and, finally, the threshold $b_0$ was calculated from $\gamma$. With the comparator, the sampled ED output was quantized to bits 0 and 1 using the obtained value of $b_0$ as the threshold. To check for packet errors, the received bits in each packet were compared with the intended transmit data.

In the case of PPM, the sampled ED output in Fig. 3.12(a) is fed to a PPM decoder which operates with the principle which was depicted in Fig. 3.6. The PPM decoder stores $2^B$ ED output samples during one symbol and decides that the slot with the highest amplitude is the “on” slot and the other ones “off” slots. To check for packet errors, the received symbols in each packet were compared with the intended transmit symbols. With DPPM, the sampled ED output is fed to a DPPM PL-SDD decoder which acts in a similar manner on packet-level as was depicted in Fig. 3.9. The PL-SDD decoder stores $L_{p,max}$ ED output samples and decides that the $N_0$ slots with the highest amplitudes are the “on” slots and the other ones “off” slots. $L_{p,max}$ refers to the maximum length of a DPPM packet, given by (3.17). DPPM packet errors were detected by checking if the receiver interpreted the correct slots as the “on” slots.

The simulated BFSK receiver is shown in Fig. 3.12(b). It consists of two envelope detecting signal paths. Carrier tones of $f_c = 434$ MHz and $f_{c2} = 444$ MHz are used here\(^7\). BPF and BPF2 are filters matched to the transmit tones $f_c$ and $f_{c2}$, respectively, and BPF is the same as the BPF in Fig. 3.12(a). In a similar fashion to the FIR filter coefficients of BPF, the coefficients of BPF2 are a sine wave with the flat envelope and length of 1000 samples but with frequency $f_{c2}$. The rectifiers and integrators are identical to the ones that are used in Fig. 3.12(a). Here too, the sampled ED outputs are Rayleigh- and Rician-distributed as determined by the tone of the BFSK-modulated RF input signal. If the tone is $f_c$, the signal passes through BPF and the sampled upper and lower ED outputs are Rician- and Rayleigh-distributed, respectively. If the tone is $f_{c2}$, the signal passes through BPF2 and the sampled upper and lower ED outputs are Rayleigh- and Rician-distributed, respectively. The received bit is decided by comparing the sampled ED outputs with each other. Packet errors were detected by comparing the received bits with the transmit bits.

Fig. 3.12(c) shows the simulated BPSK receiver. It uses the same BPF and

\(^7\)The frequency separation between $f_c$ and $f_{c2}$ is $f_{BB}$, i.e. equal to the data rate, but these tones are not exact multiples of $f_{BB}$. However, this did not seem to cause noticeable degradation in error performance in the simulations.
integrator as the OOK/PPM/DPPM and BFSK receivers. The LO frequency is $f_c = 434$ MHz and the LO signal phase is matched to the BPF output. The bandpass-filtered RF signal and synchronized LO output are fed to the mixer. Without noise, the mixer output is ideally non-negative when the data bit in the input signal is 1, and non-positive when the bit is 0. The mixer output is integrated with an integration time interval of 23 samples, i.e. the length of one RF cycle, and the integrator output is sampled with the sampling interval of $T_{BB}$. The sampler output is compared with 0 to decide whether the received bit is 0 or 1. Note that in contrast to the noncoherent OOK, PPM, DPPM and BFSK receivers, the sampled integrator output is normally distributed. The mean value is positive and negative with input bits 1 and 0, respectively.

3.3.2 Waveform-level PER simulation results

Fig. 3.13 shows the waveform-level PER simulation results of [I] and the theoretical PERs against $\gamma$. The simulated results match well with the theory. This suggests that the PER equations and calculations are correct in this work and [I]. With each modulation scheme and each level of $\gamma$ used in the simulations, transmission of as many packets was simulated as was required for at least 100 packet errors to occur. A horizontal line is used to denote the PER target used in this work, $4.8 \cdot 10^{-4}$ with 48-bit packets, which according to (3.24) is the PER that corresponds to $BER = 10^{-5}$. As the horizontal axis is SNR $\gamma$, assuming equal noise bandwidths, the figure depicts the differences in the peak output powers required with the compared modulation schemes for achieving equal error probability [I]. The information about these differences is utilized in the energy efficiency comparison in Chapter 4.

The theoretical PERs with OOK, BPSK and BFSK in the figure are obtained with (2.3), (2.7) and (2.8) by substituting the results to (3.24). The theoretical
PERs with PPM can be calculated from the SERs given by (3.7) with the equation

\[ \text{PER} = 1 - (1 - \text{SER})^{N_s}, \]

where \( N_s \) is the number of symbols required for a 48-bit packet. With 16-PPM and 64-PPM, the numbers of required symbols are 12 and 8, respectively. The PERs with the DPPM schemes are given by (3.23) that can be evaluated, for example, by performing the integration numerically in MATLAB. Calculating the integral requires the values for \( N_O \) and \( N_Z \), i.e. the number of “on” slots and the maximum number of “off” slots. With 16-DPPM, \( B = 4 \) bits are encoded per symbol and \( N_s = 12 \) symbols are required per packet. Thus, using (3.14) and (3.19), we get \( N_O = 13 \) and \( N_Z = 192 \). With 64-DPPM, the corresponding values are \( B = 6, N_s = 8, N_O = 9 \) and \( N_Z = 512 \).

It is notable that the \( \gamma \) values for the simulated points in Fig. 3.13 are actual non-scaled values obtained from the simulated signals and calculated as \( \gamma = E_{\text{sig}}/E_n \) [I]. \( E_{\text{sig}} \) is the energy conveyed by the unmodulated RF carrier over a time interval whose length is one baseband clock cycle \( T_{BB} \). \( E_n \) is the energy of the AWG noise in the expected noise bandwidth \( f_{BB} \) over the time interval \( T_{BB} \). The energies were calculated as follows. Let \( s \) and \( n \) be the modulated RF signal vector without AWG noise and the AWG noise vector, respectively, each with the length of a 48-bit packet. The packet length is \( N_{\text{tot}} \) baseband clock periods and \( N_{\text{tot}} \) depends on the modulation scheme.\(^8\) \( E_{\text{sig}} \) was calculated as

\[ E_{\text{sig}} = \frac{1}{N_{\text{act}}} \sum_{i=1}^{L} s_i^2, \]

where \( i \) is the sample index and \( L \) [samples] is the length of \( s \). \( N_{\text{act}} \) is the number of baseband clock cycles in signal \( s \) where the carrier is present. \( N_{\text{act}} \) is 48 in the case of BPSK and BFSK because, using them, the carrier signal is transmitted continuously during each baseband clock cycle. With OOK, \( N_{\text{act}} \) is the number of bits with the value 1 in the packet and it is 24 on average with equiprobable bits. With 16-PPM and 64-PPM, the value of \( N_{\text{act}} \) is 12 and 8, respectively, as a packet contains one “on” slot per symbol. With DPPM, the value of \( N_{\text{act}} \) is determined by the number of “on” slots per packet, i.e. it is 13 and 9 with 16-DPPM and 64-DPPM, respectively. Let \( N \) be the single-sided amplitude spectrum\(^9\) calculated from the noise signal \( n \). \( E_n \) was calculated as

\[ E_n = \frac{1}{N_{\text{tot}}} \sum_{i=a}^{b} N_i^2, \]

\(^8\)With OOK, BPSK and BFSK, \( N_{\text{tot}} \) is 48, i.e. the number of bits per packet. With the PPM schemes, \( N_{\text{tot}} \) is \( N_s \cdot 2^B \). With 16-DPPM and 64-DPPM, \( N_{\text{tot}} \) is equal to \( L_{p,\text{max}} \), obtained using (3.17).

\(^9\)Single-sided spectrum here refers to the spectrum after a two-sided spectrum is converted to single-sided. The single-sided spectrum contains the same amount of energy as the time-domain signal vector which the spectrum is calculated from.
where \( i \) is the index of a bin and \( a \) and \( b \) are the indices of the first and last bin inside the noise band.

During the simulations, also a second estimate of \( \gamma \) was calculated based on the observed sampler output distributions. As expected, the sampler output is Rayleigh- or Rician-distributed with the noncoherent OOK, BFSK, PPM and DPPM receivers depending on the presence and absence of the RF signal at the envelope detector input. With the BPSK receiver, the sampler output is normally distributed but the mean value depends on the data bit. Estimated values of \( A \) and \( \sigma \), denoted as \( A_{est} \) and \( \sigma_{est} \), were calculated from the sampler output distributions using the \texttt{mle()} function of MATLAB, and this second \( \gamma \) estimate was calculated as \( \gamma_{est} = A_{est}^2 / 2\sigma_{est}^2 \). These values of \( \gamma_{est} \) matched well with the values of \( \gamma \) that were calculated from the signal and noise energies as \( \gamma = E_{sig} / E_n \). This suggests that the receivers work well – the SNR in the unquantized receiver output is equal to that of the RF signal. Moreover, this suggests that \( A \) and \( \sigma \) are related to the signal and noise energies as \( \gamma = A^2 / 2\sigma^2 = E_{sig} / E_n \).

### 3.3.3 Simulation remarks and conclusions

This simulation study and the results suggest that the packet error probability is practically determined by the amount of the received signal energy. According to [7], this is expected when a matched filter or a sufficient approximation of it is used and the receiver input contains the signal and white noise. The energy-dependency of the error probability can be reasoned as follows. For the simulated points in Fig. 3.13, the \( \gamma \) values on the horizontal axis were obtained as \( \gamma = E_{sig} / E_n \). \( E_{sig} \) is the amount of energy the carrier wave conveys during a baseband clock cycle \( T_{BB} \), calculated here using (3.26), and \( E_n \) the energy of the emulated noise signal \( n \) in the expected noise bandwidth \( f_{BB} \) over the time interval \( T_{BB} \), given by (3.27). This noise energy \( E_n \) at a receiver can generally be expressed as \( E_n = P_n \cdot T_{BB} \). By substituting the \( P_n \) of (2.23), this noise energy becomes \( E_n = k \cdot T \cdot BW_N \cdot T_{BB} \). Therefore, \( \gamma = E_{sig} / E_n \) can be written as

\[
\gamma = \frac{E_{sig}}{k \cdot T \cdot BW_N \cdot T_{BB}}. \tag{3.28}
\]

Note that the noise within the presumed noise bandwidth \( BW_N = f_{BB} \) was used to calculate \( E_n \) when the \( \gamma \) of the emulated RF signals were calculated. Because the simulated PERs matched well with the theory, this implies that the noise bandwidth is \( BW_N = f_{BB} \). As \( f_{BB} = 1/T_{BB} \), we have \( BW_N = 1/T_{BB} \). Substituting this \( BW_N \) to (3.28) cancels out \( BW_N \) and \( T_{BB} \) in the denominator and yields

\[
\gamma = \frac{E_{sig}}{k \cdot T}. \tag{3.29}
\]

The Boltzmann constant \( k \) in (3.29) is a fixed value and the noise temperature \( T \) may also be assumed to have a constant value which does not depend on the modulation scheme. Thus, it can be deduced that it is the received carrier energy \( E_{sig} \) which practically determines the SNR \( \gamma \) that furthermore determines the
PER. As mentioned, this energy-dependency of the error probability has been discussed in [7]. Note that there is a connection to the discussion related to the dilation property of Fourier transform in Section 2.3. It was concluded there that, for instance, a tenfold increment of $f_{BB}$, i.e. a tenfold decrement of $T_{BB}$, reduces the SNR tenfold due to the increased noise bandwidth. A similar effect is seen in (3.29): a tenfold decrement of $T_{BB}$ reduces $E_{sig}$ and, consequently, also the SNR $\gamma$ in (3.29) both tenfold.

The above consideration is partly based on the receiver simulations that considered OOK, BPSK, BFSK, PPM and DPPM. The conclusions here can not be directly generalized to all modulation schemes. The author suspects that the result applies to schemes that utilize similar waveforms and are received using similar receivers and reception principles which were used here. The signal-energy-dependency of the error probability may apply, for instance, to M-PSK, M-FSK and QAM. However, further study would be required to show that and such analysis is considered out of the scope of this work. Nonetheless, regarding OOK, BPSK, BFSK, PPM and DPPM transmitters, this consideration has clear implications. Firstly, it implies that the uplink capability of a transmitter significantly depends on the symbol energy in the TX output signal. The output signal power and noise bandwidth each alone have a reduced meaning. The second implication is, consequently, that the benefit of using a modulation scheme that requires a low SNR per bit is that a given PER (or BER) is achieved with less received signal energy per bit, i.e. with less transmitted energy per bit. With a given modulation scheme, it ideally does not matter if the signal energy is transmitted using a short high-amplitude waveform or a long low-amplitude waveform. Equal error probability is achievable with each waveform if their energies are equal.

It was stated in the previous considerations that the values of $\gamma$ for the simulated points in Fig. 3.13 were obtained as $E_{sig}/E_n$ with $E_{sig}$ and $E_n$ calculated using (3.26) and (3.27), respectively. In the calculation of the noise energy $E_n$, an assumed noise bandwidth of $f_{BB}$ was used. Using this noise bandwidth, the simulated data in Fig. 3.13 matches well with the theoretical packet error ratios. Hereby, as deduced in [I], the simulation results suggest that the noise bandwidth is $f_{BB}$ with all the modulation schemes whose PER was simulated here. As discussed in [I], this statement is supported by [17] which, referring to ASK\textsuperscript{10}, PSK and FSK, states that all digital radio links require a receiver noise bandwidth that is equal to the symbol rate. With OOK, BPSK and BFSK, the symbol rate is equal to the data rate $R_b$ which is, furthermore, equal to $f_{BB}$. Thus, the noise bandwidth is $f_{BB}$ with them. As OOK, PPM and DPPM signals are similar, it is logical that the noise bandwidth with PPM and DPPM is the same as with OOK, $f_{BB}$, although it is not the symbol rate with them [I].

Related to the energy efficiency FOM calculations in this thesis, it can be pointed out about signal-to-noise ratio that the SNR $\gamma$ hereby refers to the ratio between the energy of the carrier and the energy of the noise in the noise

\textsuperscript{10}OOK is one form of ASK.
bandwidth, each over the time interval $T_{BB}$. In the link strength calculations, we evaluate the maximum SNR achievable with the transmitted output signal, $\gamma_{\text{max}}$. It can be expressed as

$$\gamma_{\text{max}} = \frac{E_{\text{sig, out}}}{E_n}, \quad (3.30)$$

where $E_{\text{sig, out}}$ is the amount of energy conveyed by the output carrier wave over the duration $T_{BB}$. In the FOM calculations, $\gamma_{\text{req}}$ is correspondingly the $\gamma$ required by the utilized modulation scheme and $\gamma_{\text{req}}$ practically refers to a signal-energy-to-noise-energy ratio required by the scheme. Assuming a constant signal envelope, (3.30) can be rewritten using output power and noise power as

$$\gamma_{\text{max}} = \frac{P_{\text{out}} \cdot T_{BB}}{P_n \cdot T_{BB}} \quad (3.31)$$

and finally as

$$\gamma_{\text{max}} = \frac{P_{\text{out}}}{P_n}, \quad (3.32)$$

where $P_{\text{out}}$ is hereby the peak output power and $P_n = k \cdot T \cdot BW_N$ the noise power in the noise bandwidth. $\gamma_{\text{max}}$ is thus determined by the peak output power of a transmitter as opposed to average output power. Due to the inherent duty-cycling, the average output powers of PPM and DPPM transmitters can be very low [I], [II]. However, due to the above, the low average output power does not mean that the link strength would be low – it is rather the peak output power as opposed to the average output power that, together with noise power and $\gamma_{\text{req}}$, determines the link strength.

It is to be noted that a square envelope was used with all the modulation schemes in the simulations which were presented here and, correspondingly, the receivers used bandpass filters whose coefficients were a carrier sine wave the length of $T_{BB}$ with a flat envelope. The noise bandwidth may be different if another kind of bandpass filter is used in the receiver than that used here.

### 3.4 Summary

Conventionally, ULP narrowband transmitters have used binary modulation schemes yet the use of M-ary modulation offers reduction of consumed energy per bit. This is due to the decreased active time per bit of the high-power RF circuits. Any resulting improvement in energy efficiency, however, may generally require the use of an energy-efficient modulation scheme that enables low active TX power consumption. In this section, two prospective modulations for ULP transmitters were considered: PPM and DPPM.

M-ary PPM is an orthogonal modulation and, for this, it requires a low SNR per bit to achieve a given error probability. This enables reduced PA energy consumption per bit. With PPM, when the RF signal is generated to represent
an “on” slot, the transmitted power must be higher than with BPSK. However, the RF signal is generated for a shorter duration per bit when multiple bits are encoded per symbol. The PPM signal resembles an OOK signal and can be generated using a low-complexity low-power RF front-end. The modulator can be implemented as a counter-based low-power digital block. Moreover, increasing $B$ requires only rather simple changes in the circuitry – the increment is achieved simply by increasing the word length in the digital modulator block. These properties enable low active power consumption for PPM transmitters. Therefore M-ary PPM schemes are considerable options for ULP transmitters.

M-ary DPPM offers similar benefits as PPM. It is not orthogonal but, on packet-level, offers decoding methods similar to those used for decoding an orthogonal PPM signal: instead of looking for one peak in the ED output within the duration of one symbol, the receiver may look for multiple peaks within the duration of a packet. A packet-level soft-decision decoding scheme was presented. With this PL-SDD scheme and sufficiently high $B$, a lower SNR per bit is required to achieve certain PER compared to coherent BPSK, noncoherent OOK and noncoherent BFSK. Thus, in a similar fashion to PPM, DPPM offers reduced energy consumption per bit. Similarly, DPPM supports low-complexity low-power RF front-ends and the modulation can be performed using a counter-based low-power digital block.

The energy efficiency of PPM and DPPM comes at the cost of data rate which reduces quickly as $B$ is increased. Another shortcoming compared to OOK, BPSK and BFSK is that, at the receiver side, the PL-SDD reception requires increased complexity on the digital side – memory may be required to store all the envelope detector output samples in a full-sized packet and digital signal processing is required for decoding. However, the amount of required memory could perhaps be reduced – after all, the receiver is only interested in the amplitude and time of occurrence of a reduced amount of highest peaks in the envelope detector output. The receiver optimization is out of the scope of this thesis. For lesser complexity, the receiver could also use hard-decision decoding (i.e. compare envelope detector output to a threshold to decide whether the slot is an “on” or “off” slot) but this is expected to degrade the error performance. Finally, DPPM reception is prone to insertion and deletion errors that can corrupt data in multiple symbols. This can be counteracted by the use of packet-mode transmission which isolates any errors to the contents of a single packet.

In the next chapter, we shall consider how the choice of modulation scheme affects the combined energy consumption of the two most power-consuming blocks in ULP transmitters: the LO and PA. The results of the analysis support the claim that the use of PPM and DPPM could be more energy-efficient compared to the use of OOK, BPSK and BFSK.
4. Circuit-related energy efficiency comparison of modulation schemes

4.1 Background information

This chapter discusses the method which was used in [I] for comparing the energy efficiencies of modulation schemes in a new way. The energy efficiencies of modulation schemes are traditionally compared by the SNR they require per bit to achieve a certain error probability, i.e. by comparing the performance such as those shown in Fig. 2.5. The more SNR the modulation scheme requires per bit, the more energy the PA is expected to consume per bit. This SNR requirement is a good guideline for choosing an energy-efficient modulation scheme for a transmitter if the PA can be assumed to dominate the total power consumption. However, in a ULP transmitter, the output power and PA power consumption can be very low. Consequently, other blocks, particularly the carrier synthesizer [11], can have a significant impact on the EPB. Therefore, the SNR requirement of the modulation scheme may have a less significant impact on the overall transmitter power consumption and energy efficiency. For these reasons, the energy efficiency comparison method was presented in [I] which considers how the choice of modulation scheme impacts, at the transmitter side, the combined energy consumption of the PA and the carrier synthesizer. Note that the carrier synthesizer is denoted in this discussion simply as LO although, instead of a bare LO, it could also be some other circuit such as a PLL.

In this section, a simple example comparison is provided to introduce the comparison method. Section 4.2 reviews the derivation of the equations which were used for the comparison in [I]. In [I], 16-ary and 64-ary PPM and DPPM schemes were compared with the modulation schemes conventionally used in ULP transmitters, i.e. BPSK, OOK and BFSK. Section 4.3 discusses the obtained comparison results. The results of [I] bring about a new view of the energy efficiency of the modulation schemes: if low output power is required for a radio link and the LO is expected to dominate the power consumption instead of the PA, the active ratio has a significant impact on the energy consumed per transmitted bit. The results suggest that the M-ary PPM and DPPM schemes
are more energy-efficient than OOK, BPSK and BFSK [I].

As the transmitter presented in [I] transmits 48-bit packets using DPPM, the comparison in [I] was made considering the packet error performance with 48-bit packets and using \( \text{PER} = 4.8 \times 10^{-4} \) as the targeted error probability. It can be calculated with (3.24) that this PER is equivalent to \( \text{BER} = 10^{-5} \), the BER requirement commonly used to benchmark modulation schemes in literature, when \( N_b = 48 \) bits are transmitted per packet. The PERs of the chosen modulation schemes with this packet size are plotted in Fig. 3.13. The comparison was performed with equal \( T_{BB} \) and \( f_{BB} \). With equal \( f_{BB} \), the noise bandwidths are equal with the modulation schemes discussed here. Therefore, the differences in \( \gamma \) at the crossings of \( \text{PER} = 4.8 \times 10^{-4} \) in Fig. 3.13 indicate the relative TX output powers required by these modulation schemes. For example, OOK requires 6.5 dB greater \( \gamma \) than BPSK. Therefore, if OOK is used instead of BPSK, the peak output power must be multiplied by a factor of \( 10^{6.5/10} \approx 4.47 \).

Fig. 4.1 shows the concept behind the comparison. The figure depicts the power consumption profiles of BPSK, OOK and 64-PPM transmitters during transmission of six bits. The BPSK TX is continuously active and transmits six bits over a time interval of \( 6 \cdot T_{BB} \). In this example, the LO and PA powers of the BPSK transmitter are equal. However, the analysis has been extended in [I] to take into account various ratios between the LO and PA power consumption. To achieve equal PER with the BPSK transmitter, the output powers of the OOK and 64-PPM transmitters have been scaled according to Fig. 3.13. With OOK, six bits are transmitted in \( 6 \cdot T_{BB} \) in a similar fashion to BPSK. As per Fig. 3.13, OOK requires 6.5 dB more carrier power than BPSK to achieve the targeted PER. Let us denote this output power scaling factor as \( G_{P,\text{OOK}} = 10^{6.5/10} \approx 4.47 \). For simplicity, we assume that the PA efficiency does not depend on the output power. The PA of the OOK transmitter thus consumes 4.47 times the power consumed by the PA of the BPSK transmitter. On average, half of the transmitted bits are ones and the figure depicts the transmission of an on-off keyed bit sequence 101010 \(_2\). The 64-PPM part of the figure only shows one “on” slot because six bits are encoded per 64-PPM symbol and conveyed with a single RF waveform the
Circuit-related energy efficiency comparison of modulation schemes

Table 4.1. EPBs of BPSK, OOK and 64-PPM transmitters in the example scenario

<table>
<thead>
<tr>
<th>mod</th>
<th>EPB (J/bit)</th>
<th>$EPB_{rel,mod}$ (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>BPSK</td>
<td>2</td>
<td>100</td>
</tr>
<tr>
<td>OOK</td>
<td>2.74</td>
<td>137</td>
</tr>
<tr>
<td>64-PPM</td>
<td>0.65</td>
<td>32.5</td>
</tr>
</tbody>
</table>

length of $T_{BB}$. According to Fig. 3.13, 64-PPM (i.e. PPM with $B = 6$) requires 4.6 dB greater signal power than BPSK for the targeted PER and the output power is scaled by factor $G_{P,64-PPM} = 10^{4.6/10} \approx 2.88$. 64-PPM encodes six bits per symbol and therefore the LO and PA are active for one $T_{BB}$ only.

Based on the above information, the energy consumptions of these example BPSK, OOK and 64-PPM transmitters can be calculated and compared. For simplicity, let the grid units of the time and power axes in Fig. 4.1 be 1 s and 1 W, respectively. The LO and PA of the BPSK TX each consume 1 W and the TX is active for 6 s. The energy consumed for the transmission of the six bits is therefore $2 \times 6 = 12$ J and the EPB is $EPB_{BPSK} = 12$ J/6 bits = $2$ J/bit. Similar calculations for the OOK TX yield a total consumed energy of $(1 + 4.47) \times 3 = 16.41$ J and $EPB_{OOK} = 16.41$ J/6 bits $\approx 2.74$ J/bit. For the 64-PPM TX, we get a total consumed energy of $(1 + 2.88) \times 1 = 3.88$ J and $EPB_{64-PPM} = 3.88$ J/6 bits $\approx 0.65$ J/bit.

The energy efficiency comparison is performed by considering the ratio between the energies consumed per bit using modulation scheme $mod$ and a reference modulation scheme $ref$. As per [I], the equation for the comparison is

$$EPB_{rel,mod} = \frac{EPB_{mod}}{EPB_{ref}}.$$  \hspace{1cm} (4.1)

BPSK was arbitrarily chosen as the reference modulation scheme in [I]. Thus, it is used as the reference here as well. Table 4.1 sums up the results regarding the example BPSK, OOK and 64-PPM transmitters. Using (4.1), the relative EPB with OOK has been calculated as

$$EPB_{rel,OOK} = \frac{2.74}{2} = 137\%.$$  \hspace{1cm} (4.2)

Similar calculation for 64-PPM results in $EPB_{rel,64-PPM} = (0.65)/(2) = 32.5\%$. The use of 64-PPM enables the lowest LO and PA energy consumption per bit in this example case. It consumes 68\% (4.9 dB) and 76\% (6.3 dB) less energy per bit than BPSK and OOK, respectively. The EPB with OOK is only 37\% higher compared to BPSK. It is to be noted that OOK demodulation requires 3.5 dB more (i.e. 124\% more) SNR per bit than BPSK but here the energy consumption is not this much worse. This shows how, in terms of energy consumption, OOK in some cases may not be as suboptimal a choice as the SNR-per-bit requirement alone suggests. When the PA does not dominate the TX power consumption, the optimum modulation scheme for a low-output-power
TX is not found by observing solely the SNR per bit required by the modulation schemes [I]. For this, it is worthwhile to assess how the choice of modulation scheme affects the energy consumed per bit by blocks other than the PA.

The result of this comparison is highly dependent on the LO and PA power consumptions of the reference transmitter. In this example scenario, they were balanced. However, in an extreme case, the PA power can be significantly higher or significantly lower than the power consumed by the LO. In [I], an equation for \( EPB_{\text{rel},\text{mod}} \) was used that also accounts for this. Furthermore, a second equation was used in [I] for considering a scenario where the LO power consumption needs to be scaled when the output power is scaled. The next section reviews the derivation of these equations. Using the equations, PPM and DPPM are compared with OOK, BPSK and BFSK in Section 4.3.

### 4.2 Derivation of equations for the energy efficiency comparison

Adhering to the derivation provided in [I], the equations for the energy efficiency comparison are obtained as follows. In this comparison, we shall take into account the energy consumed by the LO and PA because they can be expected to consume the most power. ULP transmitters can be implemented without other blocks that consume significant power [10] and, thus, the power consumption of other TX blocks is assumed negligible. It is assumed that the duty-cycling of the LO and PA with OOK, PPM and DPPM can be performed ideally with no overhead time\(^1\) required for switching the blocks on or off. Furthermore, it is assumed that their standby power is negligible. Finally, any energy required for initiating the transmission of a packet is generally neglected. However, an exception regarding this is done in the case of 16-DPPM and 64-DPPM – the start pulse which starts a DPPM packet is accounted for. It is taken into account in the DPPM PER performance (see Section 3.2.4 and Fig. 3.13). Moreover, it is considered in the active ratio with DPPM.

With modulation scheme \( \text{mod} \), the total power consumption of the LO and PA during transmission of the RF carrier is

\[
P_{RF} = P_{LO} + P_{PA,\text{mod}},
\]

where \( P_{LO} \) and \( P_{PA,\text{mod}} \) are the power consumptions of the LO and PA, respectively. Taking into account the active ratio of a modulation scheme which was introduced in the end of Section 2.1.1, the LO and PA are active for \( R_{t,\text{mod}} \cdot T_{BB} \) per bit. The EPB with the reference modulation can be written as

\[
EPB_{\text{ref}} = P_{RF} \cdot R_{t,\text{ref}} \cdot T_{BB} = (P_{LO} + P_{PA,\text{ref}}) \cdot R_{t,\text{ref}} \cdot T_{BB}.
\]

Similarly, with modulation scheme \( \text{mod} \) that is compared with the reference

---

\(^1\)The overhead time can be very short, for instance, if a calibrated free-running DCO or VCO is utilized as the LO.
scheme, the EPB is

\[ EPB_{\text{mod}} = (P_{\text{LO}} + P_{P\text{A,mod}}) \cdot R_{t,\text{mod}} \cdot T_{BB}. \] (4.5)

Here, \( P_{P\text{A,mod}} \) is the power consumption of the PA after output power has been scaled with the factor \( G_{P,\text{mod}} \) for equal error performance with the reference modulation. With the scaling factor, the PA power consumption with modulation scheme \( \text{mod} \) is

\[ P_{P\text{A,mod}} = G_{P,\text{mod}} \cdot P_{P\text{A,ref}}, \] (4.6)

where \( P_{P\text{A,ref}} \) is the PA power consumption of the reference TX that uses the reference modulation. Substituting (4.6) to (4.5) yields

\[ EPB_{\text{mod}} = (P_{\text{LO}} + G_{P,\text{mod}} \cdot P_{P\text{A,ref}}) \cdot R_{t,\text{mod}} \cdot T_{BB}. \] (4.7)

Substituting (4.4) and (4.7) to (4.1) yields an equation for the EPB with modulation scheme \( \text{mod} \) relative to the reference scheme

\[ EPB_{\text{rel,mod}} = \frac{(P_{\text{LO}} + G_{P,\text{mod}} \cdot P_{P\text{A,ref}}) \cdot R_{t,\text{mod}} \cdot T_{BB}}{(P_{\text{LO}} + P_{P\text{A,ref}}) \cdot R_{t,\text{ref}} \cdot T_{BB}}. \] (4.8)

To take into account various ratios between the PA and LO power consumptions, let us denote the PA power with the reference scheme as a function of the LO power. This is written as

\[ P_{P\text{A,ref}} = \alpha \cdot P_{\text{LO}}, \] (4.9)

where \( \alpha \) is a variable that depicts the ratio between the power consumptions of the PA and LO of the reference TX. To take into account factor \( \alpha \), (4.9) is substituted to (4.8) which yields

\[ EPB_{\text{rel,mod}} = \frac{(P_{\text{LO}} + G_{P,\text{mod}} \cdot \alpha \cdot P_{\text{LO}}) \cdot R_{t,\text{mod}} \cdot T_{BB}}{(P_{\text{LO}} + \alpha \cdot P_{\text{LO}}) \cdot R_{t,\text{ref}} \cdot T_{BB}}. \] (4.10)

Removing the common terms \( P_{\text{LO}} \) and \( T_{BB} \) from the numerator and denominator simplifies the equation and yields

\[ EPB_{\text{rel,mod}} = \frac{(1 + \alpha \cdot G_{P,\text{mod}}) \cdot R_{t,\text{mod}}}{(1 + \alpha) \cdot R_{t,\text{ref}}}. \] (4.11)

This equation, originally presented in [I], enables energy efficiency comparison of modulation schemes with the assumption that the LO and PA dominate the energy consumption.

In the derivation of (4.11), it is assumed that the LO power consumption does not need to be scaled when the TX output power is scaled. However, as discussed in [I], scaling the output power could, in some cases, require changes or allow for changes in the LO power consumption – increased output power could lead to an increased LO power consumption and reduced output power could allow for a reduced LO power consumption. For this, the second comparison equation was
derived in [I] for a scenario where $P_{LO}$ scales with the PA power consumption. For simplicity, such a scenario was considered where the LO power consumption scales linearly with the output power and PA power consumption. In this scenario, the LO power consumption with modulation $mod$ is $G_{P,mod} \cdot P_{LO}$ and the EPB with modulation $mod$ given by (4.7) can be rewritten as

$$EPB'_{mod} = G_{P,mod} \cdot (P_{LO} + P_{PA,ref}) \cdot R_{t,mod} \cdot T_{BB}. \quad (4.12)$$

Substituting (4.12) and (4.4) to (4.1) now yields

$$EPB'_{rel,mod} = \frac{EPB'_{mod}}{EPB_{ref}} = \frac{G_{P,mod} \cdot (P_{LO} + P_{PA,ref}) \cdot R_{t,mod} \cdot T_{BB}}{(P_{LO} + P_{PA,ref}) \cdot R_{t,ref} \cdot T_{BB}}. \quad (4.13)$$

Removing the common terms from the numerator and denominator results in

$$EPB'_{rel,mod} = \frac{G_{P,mod} \cdot R_{t,mod}}{R_{t,ref}}. \quad (4.14)$$

This equation, originally presented in [I], enables comparison of modulation schemes in the scenario where the LO power consumption scales linearly with the PA power consumption and output power. In this case, the relative energy efficiency is a constant determined by the active ratios of the compared modulation schemes and the output power scaling factor determined by the SNR requirements of the modulation schemes.

Before applying the derived equations, it is to be mentioned that it was chosen in [I] that all the modulation schemes are used with equal $f_{BB}$ and $T_{BB}$ for equal noise bandwidths. In the case of OOK, BPSK, PPM and DPPM, this means that also the null-to-null bandwidths are equal. According to [6], the null-to-null bandwidth with OOK and BPSK is $2 \cdot R_b$, i.e. $2 \cdot f_{BB}$. It was confirmed using MATLAB that the null-to-null bandwidth is $2 \cdot f_{BB}$ also with PPM and DPPM [I]. However, with noncoherent BFSK, the null-to-null bandwidth is wider, $3 \cdot R_b$ [6], i.e. $3 \cdot f_{BB}$, assuming BFSK frequency separation of $R_b$ which is required for orthogonality. Because of this, we shall consider two BFSK schemes in the comparison: scheme BFSK$_1$ with the same $f_{BB}$ and $T_{BB}$ with the other modulation schemes but a wider null-to-null bandwidth, and scheme BFSK$_2$ with $f_{BB}$ and the frequency separation $\Delta f$ scaled down for equal null-to-null bandwidth with the other modulation schemes.

### 4.3 Circuit-related energy efficiency comparison of OOK, BPSK, BFSK, PPM and DPPM

This section discusses the results of the energy efficiency comparison in [I], performed using the method described in the previous section and [I]. Table 4.2 lists the parameters used in the calculations. The second column lists the $\gamma \quad (4.14)$

\[\text{This is also supported by the measured spectra with DPPM signals (see Fig. 5.6).}\]
Table 4.2. Parameters for energy efficiency comparison calculations with BPSK as the reference modulation

<table>
<thead>
<tr>
<th>mod</th>
<th>$\gamma_{req}$ (dB)</th>
<th>$\Delta\gamma_{req}$ (dB)</th>
<th>$G_{P,mod}$</th>
<th>$R_{t,mod}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>BPSK</td>
<td>9.6</td>
<td></td>
<td>1.0</td>
<td>1</td>
</tr>
<tr>
<td>OOK</td>
<td>16.1</td>
<td>6.5</td>
<td>4.47</td>
<td>1/2</td>
</tr>
<tr>
<td>BFSK$_1^{(1)}$</td>
<td>13.4</td>
<td>3.8</td>
<td>2.40</td>
<td>1</td>
</tr>
<tr>
<td>BFSK$_2^{(2)}$</td>
<td>–</td>
<td>–</td>
<td>1.60</td>
<td>1.5</td>
</tr>
<tr>
<td>16-PPM</td>
<td>13.8</td>
<td>4.2</td>
<td>2.63</td>
<td>1/4</td>
</tr>
<tr>
<td>64-PPM</td>
<td>14.2</td>
<td>4.6</td>
<td>2.88</td>
<td>1/6</td>
</tr>
<tr>
<td>16-DPPM$^{(3)}$</td>
<td>14.6</td>
<td>5.0</td>
<td>3.16</td>
<td>13/48</td>
</tr>
<tr>
<td>64-DPPM$^{(3)}$</td>
<td>14.8</td>
<td>5.2</td>
<td>3.31</td>
<td>9/48</td>
</tr>
</tbody>
</table>

Table obtained from [I], used under CC BY 4.0.

1) BFSK scheme with equal $T_{BB}$ and $f_{BB}$ but wider $BW_{null}$.
2) BFSK scheme with $f_{BB}$ and $\Delta f$ scaled for equal $BW_{null}$ with other modulations.
3) Packet-mode scheme as discussed in Section 3.2.3 with GSs and start pulse.

The modulation schemes require for PER = 4.8·10$^{-4}$, denoted as $\gamma_{req}$. The $\gamma$ requirements can be seen in Fig. 3.13. $\Delta\gamma_{req}$ is the difference in the required $\gamma$ between modulation $mod$ and the reference modulation scheme, obtained as

$$(\Delta\gamma_{req})_{dB} = (\gamma_{req,mod})_{dB} - (\gamma_{req,ref})_{dB}. \label{eq:4.15}$$

Here $\gamma_{req,mod}$ and $\gamma_{req,ref}$ are the $\gamma$ required by modulation $mod$ and the reference modulation scheme, respectively. BPSK was arbitrarily chosen as the reference modulation. $G_{P,mod} = 10^{(\Delta\gamma_{req})_{dB}/10}$ is the power scaling factor with modulation $mod$ [I]. It is assumed that modulation $mod$ achieves equal PER with the reference modulation scheme when the output power is scaled by $G_{P,mod}$. $R_{t,mod}$ is the active ratio of the modulation scheme. The active ratios with OOK, BFSK, PPM and DPPM were discussed in Sections 2.1.1, 2.1.2 and 2.1.3, respectively. The active ratio with PPM was discussed in Section 3.1 and it is $1/B$. With the packet-mode DPPM scheme discussed in Section 3.2.3, the active ratio is given by (3.20). With 16-DPPM and 64-DPPM, $B$ is 4 and 6, respectively, and the active ratios are 13/48 and 9/48, respectively.

The table exceptionally contains two rows for BFSK because of the wider null-to-null bandwidth, $BW_{null}$, which was discussed on the previous page. The null-to-null bandwidth is 3$f_{BB}$ with BFSK and 2$f_{BB}$ with OOK, BPSK, PPM and DPPM. For this, we consider two BFSK schemes, denoted as BFSK$_1$ and BFSK$_2$. BFSK$_1$ uses the same $T_{BB}$ and $f_{BB}$ as the other modulation schemes and its null-to-null bandwidth is wider but noise bandwidth is equal$^{3}$ to the other

$^{3}$The noise bandwidth is $f_{BB}$ with OOK, BPSK, BFSK, PPM and DPPM as was discussed in Section 3.3.3.
modulations. With BFSK₂, $f_{BB}$ and $\Delta f$ are scaled by a factor of 2/3 for an equal null-to-null bandwidth with the other modulation schemes, i.e. the baseband clock period $T_{BB}$ is scaled by 3/2 and it is 50% longer. For the derived energy efficiency equations to apply, the active ratio of BFSK₂ is referenced to the baseband clock period $T_{BB}$ that the other modulation schemes use. As BFSK₂ uses $T_{BB}$ extended by 50%, its active time per bit is 50% greater compared to BFSK₁ and the active ratio can be written as $R_{t,BFSK₂} = 1.5 \cdot R_{t,BFSK₁} = 1.5$. The signal bandwidth and noise bandwidth with BFSK₂ are 1/3 narrower than with BFSK₁ because of the lower $f_{BB}$. For this, the output power can be reduced⁴ by a factor of 1/3 compared to BFSK₁ and we have $G_{P,BFSK₂} = (2/3) \cdot G_{P,BFSK₁} \approx 1.60$.

The EPBs relative to BPSK, the reference modulation scheme, are obtained by substituting the $R_{t,mod}$ and $G_{P,mod}$ values of Table 4.2 to (4.11). For instance, the relative EPB with OOK is

$$EPB_{rel,OOK} = \frac{(1 + \alpha \cdot 4.47) \cdot 0.5}{(1 + \alpha) \cdot 1.0}. \quad (4.16)$$

Similarly, the relative EPB with 64-PPM is

$$EPB_{rel,64-PPM} = \frac{(1 + \alpha \cdot 2.88) \cdot 1/6}{(1 + \alpha) \cdot 1.0}. \quad (4.17)$$

The equations for the other modulation schemes can be derived in a similar fashion. The EPBs relative to BPSK are plotted in Fig. 4.2 against $\alpha$. Factor $\alpha$ here depicts the ratio of the power consumptions of the PA and LO of the reference BPSK transmitter only. Due to the power scaling, the ratios between the PA and LO power consumptions are different with the other modulation schemes. At any vertically drawn line, the values express how much the EPB

⁴See the discussion related to the dilation property of the Fourier transform in Section 2.3.
is relative to BPSK after switching the modulation scheme from BPSK to the modulation scheme under study and after scaling the output power for equal error performance. Because the output powers have been scaled for equal error performance, the relative EPBs with all the modulation schemes on any vertically drawn line can be compared with each other, not just with the reference scheme. The figure basically answers the question: if a BPSK transmitter was designed with some PA and LO power consumption whose ratio is described by \( \alpha \), how much would the EPB approximately be if modulation was changed to modulation scheme \( \text{mod} \) and output power was scaled for equal PER in the AWGN channel with the assumptions mentioned in Section 4.2.

In Fig. 4.2 on the right, \( \alpha \) is high and the PA dominates the power consumption. In this region, the energy efficiency of a modulation scheme is expectedly determined by the SNR it requires per bit for the targeted error probability. However, the relative efficiencies are significantly different in the regions where \( \alpha \) is lower than 10. It can be calculated that (4.11) approaches \( \frac{R_{t,\text{mod}}}{R_{t,\text{ref}}} \) as \( \alpha \) approaches 0. Thus, the lower the value of \( \alpha \) is, the more the energy efficiency is impacted by the active ratio. At the region where \( \alpha < 0.1 \), the LO dominates the power consumption and the energy efficiency of a modulation scheme is mostly determined by the active ratio – the less time the PA and LO are active per bit, the better. Regarding ULP transmitters, the areas outside the range from \( \alpha = 0.1 \) to \( \alpha = 10 \) may present somewhat uncommon scenarios as, generally, the PA and LO power consumptions are not extremely imbalanced. In the region between \( \alpha = 0.1 \) and \( \alpha = 10 \), the energy efficiency is affected by both the active ratio and the SNR the modulation scheme requires per bit.

The 16-ary and 64-ary PPM and DPPM schemes possess two favorable properties regarding transmitter-side energy efficiency: low required SNR per bit and low active time per bit [I]. For this, they perform well compared to the conventionally utilized binary modulation schemes in all these regions. The key implication here is that, no matter what the ratio between the LO and PA power consumptions is, use of the pulse-position schemes can be expected to conserve energy on the transmitter side compared to BPSK, OOK and BFSK [I] – the same error performance is achieved with less energy consumed per bit.

The main interest in this work and [I] is 64-DPPM. For another point of view, the relative EPBs were plotted in [I] also using 64-DPPM as the reference modulation scheme. When the reference modulation scheme is changed from BPSK to another scheme, the \( G_{P,\text{mod}} \) values of Table 4.2 do not apply because the \( \Delta \gamma_{\text{req}} \) are changed. Table 4.3 shows the parameters for the energy efficiency calculations with the values of \( \Delta \gamma_{\text{req}} \) and \( G_{P,\text{mod}} \) recalculated using 64-DPPM as the reference. The SNR requirement, \( \gamma_{\text{req}} \), and the active ratio, \( R_{t,\text{mod}} \), are properties of a modulation scheme and they are thus the same as in Table 4.2.

Fig. 4.3 shows the EPBs relative to 64-DPPM. They can be obtained using (4.11) with the values from Table 4.3. This figure answers the question: if a 64-DPPM transmitter was designed with some PA and LO power consumption whose ratio is \( \alpha \), how much would the EPB approximately be if modulation
Table 4.3. Parameters for energy efficiency comparison calculations with 64-DPPM as the reference modulation

<table>
<thead>
<tr>
<th>mod</th>
<th>$\gamma_{req}$ (dB)</th>
<th>$\Delta\gamma_{req}$ (dB)</th>
<th>$G_{P,mod}$</th>
<th>$R_{t,mod}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>BPSK</td>
<td>9.6</td>
<td>-5.2</td>
<td>0.302</td>
<td>1</td>
</tr>
<tr>
<td>OOK</td>
<td>16.1</td>
<td>1.3</td>
<td>1.35</td>
<td>1/2</td>
</tr>
<tr>
<td>BFSK$_1$</td>
<td>13.4</td>
<td>-1.4</td>
<td>0.724</td>
<td>1</td>
</tr>
<tr>
<td>BFSK$_2$</td>
<td>–</td>
<td>–</td>
<td>0.483</td>
<td>1.5</td>
</tr>
<tr>
<td>16-PPM</td>
<td>13.8</td>
<td>-1.0</td>
<td>0.794</td>
<td>1/4</td>
</tr>
<tr>
<td>64-PPM</td>
<td>14.2</td>
<td>-0.6</td>
<td>0.871</td>
<td>1/6</td>
</tr>
<tr>
<td>16-DPPM</td>
<td>14.6</td>
<td>-0.2</td>
<td>0.955</td>
<td>13/48</td>
</tr>
<tr>
<td>64-DPPM</td>
<td>14.8</td>
<td>–</td>
<td>1.00</td>
<td>9/48</td>
</tr>
</tbody>
</table>

1) BFSK scheme with equal $T_{BB}$ and $f_{BB}$ but wider $BW_{null}$.
2) BFSK scheme with $f_{BB}$ and $\Delta f$ scaled for equal $BW_{null}$ with other modulations.
3) Packet-mode scheme as discussed in Section 3.2.3 with GSs and start pulse.

scheme was changed to modulation scheme $mod$ and output power was scaled for equal PER. Here, $\alpha$ is the ratio between the PA and LO power consumptions of the reference 64-DPPM transmitter. Out of the binary modulation schemes, the nearest competitor at low $\alpha$ is OOK that consumes at least 167% more energy per bit. At high $\alpha$, the nearest competitor is BPSK that consumes at least 61% more. Considering the region between $\alpha = 0.1$ and $\alpha = 10$, using BPSK or OOK would consume 72% to 340% and 190% to 257% more energy. BFSK performs the worst here and would consume at least 291% more energy in this region. All in all, the figure suggests that using any of the conventionally used binary modulation schemes (i.e. OOK, BPSK or BFSK) would consume at least 61% more energy per bit than 64-DPPM in any scenario. It is to be noted that OOK and BFSK have been more popular in published ULP transmitters than BPSK. 64-DPPM is a particularly good choice of modulation scheme compared to them as OOK and BFSK consume at least 167% and 286% more energy per bit for the targeted error performance here.

In Fig. 4.3, there is a somewhat wide margin from 16-PPM and 16-DPPM to OOK and BFSK. Thus, also octal or even quaternary PPM and DPPM could possibly offer an improvement in energy efficiency compared to them as mentioned in [I]. The octal and quaternary schemes would allow for a higher data rate compared to the 16-ary and 64-ary schemes which were discussed here and still avoid the carrier synchronization required by coherent BPSK.

The previous calculations applied to the case where LO power consumption is assumed to require no scaling with output power. Equation (4.14) was derived for the case where the LO power consumption scales by the same amount with that of the PA. Equation (4.14) is equal to the value that (4.11) approaches as $\alpha$ approaches infinity. In this scenario, the energy efficiency of a modulation
scheme is therefore determined by the SNR it requires per bit to achieve the targeted error probability [I]. Fig. 4.4 shows the EPBs relative to BPSK in this case, calculated using (4.14) and the parameters of Table 4.2. OOK and BFSK would consume more than 2x the energy consumed with BPSK. The pulse-position schemes enable EPB reduction by 14% to 52% compared to BPSK. BPSK, OOK and BFSK consume 61%, 260% and 286% more energy per bit than 64-DPPM, respectively. Also in this scenario with the LO power consumption scaling, the pulse-position schemes are the most energy-efficient [I]. They can be expected to achieve the same error performance as the conventionally used binary modulations with less energy consumed per bit by the LO and PA.

4.4 Summary

The energy efficiencies of BPSK, OOK, BFSK, 16-PPM, 64-PPM, 16-DPPM and 64-DPPM were compared. In the comparison, the energy consumed by
the LO and PA was considered. These blocks are expected to consume the majority of power in low-complexity ULP transmitters. It was assumed that the transmitters achieve an equal error performance with these modulation schemes if their peak output powers are scaled as determined by the $\gamma$ the modulation schemes require to achieve the targeted error probability. In this work, the power-scaling factors were calculated based on the error performance in the AWGN channel. The results predict that use of PPM and DPPM could conserve energy at the transmitter side compared to use of OOK, BPSK and BFSK. In this energy efficiency comparison, two scenarios were considered: one where the LO power consumption is fixed and the other where it scales linearly with the PA power consumption.

In the scenario with fixed LO power consumption, the active ratio of a modulation scheme has significant effects. The relative energy efficiencies were plotted against $\alpha$, the ratio between the PA and LO power consumptions of a transmitter that uses a reference modulation scheme. The study suggests that, if a low $\alpha$ is expected, i.e. if the LO is expected to consume significant power compared to the PA, it is better to choose a modulation scheme with a low active ratio. In that case, OOK, PPM and DPPM enable lower energy consumption than BPSK or BFSK. If $\alpha$ is expected to be high and power consumption is more dominated by the PA, the energy efficiency of a modulation scheme is rather determined by the SNR it requires per bit. In this latter case, BPSK and the pulse-position schemes perform significantly better than OOK and BFSK. Due to a relaxed SNR-per-bit requirement and low active time per bit, PPM and DPPM perform well compared to OOK, BPSK and BFSK with any value of $\alpha$. Compared to 64-DPPM, the use of BPSK is expected to consume at least 61% more energy per bit and the use of OOK or BFSK at least 167% more with any $\alpha$ as can be seen in Fig. 4.3.

In the second scenario, the modulation schemes were compared with the assumption that the LO power consumption scales linearly with the PA power consumption. It was concluded that, in this case, the energy efficiency of a modulation scheme is determined by the SNR it requires per bit for the targeted error probability. In this scenario, OOK and BFSK perform the worst and consume more than 2x the energy consumed with BPSK and almost 4x the energy consumed with 64-DPPM. The discussed PPM and DPPM schemes enable 14% to 52% lower energy consumption compared to BPSK. Choosing BPSK, OOK or BFSK over 64-DPPM would consume 61%, 260% and 286% more energy per bit, respectively.

This section concludes the theoretical transmitter discussion of this thesis. The next chapter presents the implemented transmitter circuits including two narrowband transmitters that use DPPM. The PO-based TX achieves one of the lowest EPBs among published transmitters and, yet, enables an uplink range of 30 meters. The second TX is based on the direct-modulation architecture and it produces a higher peak output power. It achieves a state-of-the-art energy efficiency FOM and the estimated uplink range is up to 1 kilometer.
5. ULP transmitter prototypes in 180 nm CMOS

The two previous chapters discussed how the use of M-ary pulse-position modulation schemes could increase transmitter energy efficiency. This chapter presents two narrowband DPPM transmitter prototypes and a DPPM modulator for a UWB transmitter. These circuits have originally been presented in [I], [II], [III] and [IV]. The first narrowband radio transmitter prototype in [II] and the second one in [I] are based on the power oscillator and direct-modulation architectures, respectively. These architectures were discussed in Section 2.5.1 and the block diagrams are shown in Fig. 2.7. The circuits were implemented in 180 nm CMOS. The power-oscillator-based transmitter in [II] achieves one of the lowest power consumptions and EPBs among published ULP transmitters. Despite the low EPB, it was able to deliver the data over a 30-meter line-of-sight uplink in the measurements [II]. The direct-modulation transmitter of [I] does not only consume ultra-low power but is also state of the art in terms of energy efficiency, achieving an energy efficiency FOM of −189.7 dBJ/bit. Its expected uplink range is significantly greater, up to one kilometer, due to a greater SNR, enabled by higher output power and narrower noise bandwidth.

The narrowband transmitters operate in the 434-MHz band. This band offers a lower path loss compared to, for example, the 2.4-GHz industrial, scientific and medical (ISM) band. However, the use of DPPM could be equally beneficial in other frequency bands. In the UWB transmitter for which the DPPM modulator was implemented, the UWB pulse generator supports carrier frequencies from 3 to 5 GHz. All the three prototypes utilize DPPM with up to six bits encoded per symbol. The transmitters also support OOK.

5.1 Power-oscillator-based narrowband DPPM transmitter

Fig. 5.1 shows the architecture of the power-oscillator-based transmitter presented in [II]. The transmitter comprises a DPPM modulator, a multiplexer for choosing between OOK and DPPM modulation, and a power oscillator. Also a memory register (MEM REG) and control signal generator circuitry (CTRL) are included on-chip. The memory register is used to store the transmit data and
trim bits that are written through a serial peripheral interface (SPI). During data transmission, the control signal generator provides signals such as clock and data to the transmitter. The operation of the DPPM modulator is described in Section 5.1.2. Data can be transmitted both continuously and in packets. In packet-mode DPPM transmission, up to $N_s = 63$ symbols can be transmitted per packet with one to six bits encoded per symbol. The overall structure of the packet is the same as in Fig. 3.8: the packet begins with a start pulse and guard slots are utilized. OOK modulation is also supported. In OOK mode, the CTRL block feeds data bits one by one to the ENABLE input of the power oscillator, clocked at the baseband clock frequency $f_{BB}$. For this, no dedicated OOK modulator block is present. DPPM is the primary modulation due to better energy efficiency which was discussed in Chapter 4. In this prototype, the DPPM modulator and the OOK signal that is fed to the front-end are clocked using an external clock signal generator.

### 5.1.1 Power oscillator

Fig. 5.2 shows the schematic of the PO used in [II]. A current-reusing LC oscillator topology [37] is utilized. The oscillator core comprises the transistors $M_{P1}$ and $M_{N1}$, connected as a complementary cross-coupled transistor pair (XCP). The XCP is connected to an LC tank. Use of a complementary XCP reduces the power consumption compared to conventional topologies [37] such as an NMOS-NMOS XCP. OOK and DPPM RF signals can be generated by toggling the ENABLE signal. The TUNE_AMP control word can be used to coarsely control the output power. The supply voltage of the PO is 1.2 V.

The oscillation frequency is set by the LC tank. It contains 1) three digitally tunable on-chip capacitor matrices, denoted in Fig. 5.2 as $C_{coarse}$, $C_{med}$ and $C_{fine}$, and 2) an off-chip copper loop. The copper loop acts both as the inductor and the antenna. The coarse, medium and fine tuning capacitors are controlled using the 3-bit, 4-bit and 6-bit frequency control words FC3, FC2 and FC1. This PO design was also used in [III] but in a different research-oriented IC.
FC1, respectively. Each tuning bit FCX<n> controls 2^n unit capacitors in the corresponding capacitor matrix.

The coarse and medium tuning have been implemented using the tuning capacitor unit design shown in Fig. 5.3(a) [28]. It contains control switches and metal-insulator-metal (MIM) capacitors C0. According to postlayout simulations, the tuning range of the coarse-tuning capacitor array C_coarse is from 0.43 to 1.98 pF with an average step size of 222 fF. The simulated tuning range of C_med is from 0.278 to 0.519 pF with an average step size of 16 fF.

Fig. 5.3(b) shows the LSB unit of the fine tuning capacitor that is based on an idea presented in [38]. Depending on the logic value of BIT, the capacitance from the node VN to ground is either C1 or (1/C1 + 1/C2)^{-1} and the switchable capacitance is, thus, ΔC = C1 − (1/C1 + 1/C2)^{-1}. With this design, ΔC can be in the range of femtofarads. With such small switched capacitances, for instance, matching can be a challenge [38] without the use of an advanced unit capacitor design. The major disadvantage of this design is, however, a considerable amount of unswitched capacitance [38]. C1 and C2 are metal-oxide-metal (MOM)
capacitors as, with the used CMOS process, MOM capacitors can be of a smaller size than the minimum-sized MIM capacitor. For a smaller frequency tuning step, $C_{\text{fine}}$ is connected between the node VN and ground. This enables a smaller frequency tuning step compared to a case where the same amount of capacitance is toggled between the nodes VP and VN [38]. The sizes of $C_1$ and $C_2$ are 18.0 and 51.3 fF, respectively, which yields $\Delta C \approx 4.7$ fF. However, according to postlayout simulations with parasitics included, the capacitance step is 5.5 fF on average and $C_{\text{fine}}$ can be tuned from 1.07 pF to 1.41 pF.

Fig. 5.3(c) shows the on-PCB copper loop antenna, designed in Agilent Advanced Design System (ADS). A stripe width of 1 mm is used and the outer dimensions are 22 mm by 27 mm. The heights of the copper layer and the FR4 are 35 μm and 0.8 mm, respectively. No ground plane is present below the antenna as it would degrade the radiation efficiency. To avoid excessive parasitic capacitances at the nodes VP and VN, customized RF pads consisting only of metal layers are used to connect VP and VN to the antenna.

### 5.1.2 DPPM modulator

Fig. 5.1 includes a block diagram of the digital DPPM modulator that comprises an adder, a clocked counter and an equality comparator. The block generates a DPPM signal as depicted in Fig. 3.8 with the start pulse and up to 63 symbols per packet. The control signal generator, i.e. CTRL, provides the clock signal $\text{CLK}_{BB}$ and symbol $\text{SYM}$. $\text{CLK}_{BB}$ is derived from an off-chip clock signal by gating it. The number of bits per symbol, $B$, can be set from 1 to 6 by utilizing $\text{SYM}$ values limited from 0 to $2^B-1$. Thus, DPPM schemes from 2-DPPM to 64-DPPM can be used. A guard slot is added to each symbol by the adder that increments the value of $\text{SYM}$ by one.

When the transmitter is off, the counter is at the maximum value, $1111111_2$, and $\text{CLK}_{BB}$ is disabled. Transmission is initiated by enabling the clock signal. At the first rising clock edge, the counter value changes to zero. When the count is zero, the equality comparator outputs a pulse the length of $T_{BB}$ to the $\text{ENABLE}$ input of the power oscillator. This enables the oscillation and generation of the RF waveform as depicted in Fig 5.2 at the top right. Simultaneously, a new symbol is requested using the $\text{SYM\_REQ}$ signal. The counter counts up until the value is $\text{SYM}+1$ after which the count is reset to zero. Again, when the value is zero, a pulse is fed to the $\text{ENABLE}$ input of the power oscillator and a new symbol is requested. The CTRL block counts the amount of the transmitted symbols. When $N_s$ symbols have been transmitted, transmission of a packet is finished, $\text{CLK}_{BB}$ is disabled and the counter is reset to $1111111_2$.

The functionality of the DPPM modulator was written in VHDL language. The VHDL code was synthesized and the synthesized circuit was place-and-routed. A 1.2-V High-Vt digital logic library was used. The place-and-routed modulator contains 148 standard logic cells and the logic cell core area is 0.002 mm$^2$. The use of High-Vt cells enables a reduced standby power.
5.1.3 General measurement results

The PO-based transmitter was implemented in a 180 nm CMOS process and the key measurement results have been presented in [II]. A micrograph of the implemented blocks is shown in Fig. 5.4. The supply voltage of the power oscillator and the DPPM modulator was set to 1.2 V. For reduced power consumption, the amplitude tuning word of the power oscillator was set to TUNE_AMP = 001000\textsubscript{2}, less than maximum. When tuned to the frequency of 434 MHz, the PO consumes 230 μA during continuous-wave transmission. As the transmitter does not have a 50-Ω output, the output power cannot be measured directly by connecting the output to a standard 50-Ω input of a measurement device. Therefore, the radiated output power was estimated through link measurements using the measurement method explained in Section VI-A of [II]. The output power\textsuperscript{2} is approximately –25 dBm. The standby currents of the power oscillator and the DPPM modulator are 1.0 nA and 0.2 nA, respectively. The measured frequency tuning range of the power oscillator is from 515.7 to 398.6 MHz. The frequency tuning step size is lower than 120 kHz throughout this range. In a supply voltage range from 1.0 to 1.4 V, the supply sensitivity of the power oscillator is practically linear with a slope of –8.83 kHz/mV. The frequency decreases as the supply voltage is increased.

When modulated data is transmitted, the power oscillator is off part of the time. The average current consumption is therefore significantly lower compared to continuous-wave transmission. When OOK data is transmitted continuously, the average current consumption of the power oscillator and its digital control gates is 115.6 μA, 116.3 μA and 119.3 μA at data rates of 0.1 Mbps, 1.0 Mbps and 5 Mbps, respectively. The corresponding EPBs are 1.39 nJ/bit, 140 pJ/bit

\textsuperscript{2}The output power here means the power delivered to the antenna multiplied by the radiation efficiency of the antenna.
and 28.6 pJ/bit, respectively. With the higher data rates, the EPB is reduced but, as was discussed in Section 2.3, the noise bandwidth is wider, signal-to-noise ratio is lower and uplink performance is reduced.

Fig. 5.5 shows the measured current consumption, average data rate and average energy consumption per bit when DPPM data is transmitted continuously. The values are plotted against $B$, the number of bits encoded per symbol. The results are shown with baseband clock frequencies (i.e. $f_{BB}$) of 0.1, 1.0 and 5.0 MHz. The current consumption includes the current consumed by the power oscillator, its digital control gates and the DPPM modulator. With $B = 6$, the average current is 6.9, 7.3 and 8.6 $\mu$A with the baseband clock frequencies of 0.1, 1.0 and 5.0 MHz, respectively. The mean data rates of Fig. 5.5(b) can be calculated with (3.12). The EPBs of Fig. 5.5(c) have been calculated with the equation $\text{EPB} = V_{DD} \cdot I_b / R_b$. As predicted by (3.1), the EPB decreases when $B$ increases. The smallest EPB, 11.6 pJ/bit, is achieved with $B = 6$ and $f_{BB} = 5.0$ MHz in which case the average data rate is 895.5 kbps. The smallest power consumption, 8.3 $\mu$W, is achieved with $f_{BB} = 100$ kHz and $B = 6$ in which case the average data rate and EPB are 17.9 kbps and 462 pJ/bit, respectively.

It is fair to compare the performance of the transmitter in OOK and DPPM modes with equal $f_{BB}$. As was discussed in Sections 4.2 and 3.3.3, the null-to-null bandwidth and noise bandwidth are $2f_{BB}$ and $f_{BB}$, respectively, with both types of modulation. Due to equal $T_{BB}$ and peak output power, the transmitted energy in a time slot the length of $T_{BB}$ is equal with both types. Hence, the signal-to-noise ratio $\gamma$ of the signal delivered to a receiver is expected to be the same with both modulations. With $f_{BB} = 100$ kHz, the EPBs with OOK and

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![Figure 5.5](image_url)

**Figure 5.5.** (a) Current consumption, (b) average data rate and (c) average energy consumption per bit of the PO-based TX prototype as a function of $B$ (number of bits encoded per symbol) when DPPM data is transmitted continuously. Obtained from [II], used under CC BY 4.0.
DPPM are 1.39 nJ/bit and 0.462 nJ/bit, respectively. Thus, DPPM achieves a 67% lower EPB. With $f_{BB} = 5.0$ MHz, the EPBs are 28.6 and 11.6 pJ/bit and the EPB with DPPM is 59% lower. Despite the lower EPB, according to Fig. 3.13, the error performance with DPPM is slightly better. Thus, the EPB reduction does not come at the cost of error performance.

Fig. 5.6 shows example spectra of continuously transmitted OOK and DPPM signals. The baseband clock frequency $f_{BB}$ is set to 100 kHz and 5 MHz in Fig. 5.6(a) and Fig. 5.6(b), respectively. The spectra were obtained using a spectrum analyzer connected to a $\lambda/4$ whip antenna that was placed a few centimeters apart from the on-PCB transmit antenna. The measurement device noise floor and local interferers are also shown. Using DPPM, less pulses are transmitted per second and, for that, the average output power is lower compared to OOK. However, this does not imply lower uplink range nor worse error performance with DPPM because the error performance is rather determined by the $\gamma$ of the signal that is delivered to a receiver instead of the average power (see Section 3.3.3). The magnitude of this $\gamma$ depends on $\gamma_{max}$, the maximum SNR achievable with the output signal. With equal peak output power and noise bandwidths, the $\gamma_{max}$ are ideally equal with OOK and the DPPM schemes. The DPPM schemes are therefore expected to achieve a lower PER than OOK (see Fig. 3.10 and Fig. 3.13).

The spikes in the spectra are mainly local interferers and mixing products of the baseband clock at the harmonics of $f_{BB}$. An external low-jitter frequency generator was used as the baseband clock. The spikes at multiples of $f_{BB}$ are
Table 5.1. Measured transmitter performance in packet-mode with OOK and 64-DPPM

<table>
<thead>
<tr>
<th>Packet Rate</th>
<th>Data Rate</th>
<th>Average PO Duty Cycle (OOK)</th>
<th>Power Consumption (OOK)</th>
<th>Energy Per Bit (OOK)</th>
<th>PO Duty Cycle (64-DPPM)</th>
<th>Power Consumption (64-DPPM)</th>
<th>Energy Per Bit (64-DPPM)</th>
<th>EPB improvement with 64-DPPM</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>100 packets/s</td>
<td>100 kHz</td>
<td>2.4%</td>
<td>6.61 µW</td>
<td>1.38 nJ/bit</td>
<td>0.9%</td>
<td>2.48 µW</td>
<td>67.1 nW</td>
</tr>
<tr>
<td></td>
<td>4.8 kbps</td>
<td>5 MHz</td>
<td>0.048%</td>
<td>143 nW</td>
<td>29.8 pJ/bit</td>
<td>0.018%</td>
<td>67.1 nW</td>
<td>14.0 pJ/bit</td>
</tr>
</tbody>
</table>

Table obtained from [II], used under CC BY 4.0.

lowered if the baseband clock generator is, for instance, an on-chip ring oscillator whose output clock signal contains more jitter. The effect of jitter was emulated by rapidly toggling the frequency of the external frequency generator during transmission. Fig. 5.6(c) shows how the spikes are lowered in the case of OOK with \( f_{BB} = 5 \) MHz.

The performance of the transmitter was also measured in packet-mode with 48 bits encoded per packet and 100 packets transmitted per second. This equals to a fixed data rate of 4.8 kbps. With OOK, 50% of the bits were set to ones for average current consumption. With 100 packets per second and 24 bits with the value 1 per packet, 2400 pulses are transmitted per second with OOK. With DPPM, the amount of symbols per packet was \( N_s = 8 \) with \( B = 6 \) bits encoded per symbol. A start pulse was included in each packet. The DPPM symbols in each packet were randomized in such a manner [II] which resulted in the average modulator current consumption\(^3\). With DPPM, nine pulses are transmitted per packet and hence 900 pulses are transmitted per second.

Table 5.1 shows the packet-mode measurement results with OOK and DPPM with \( f_{BB} \) of 100 kHz and 5 MHz. The power oscillator duty cycles are obtained by dividing the pulse rate (the number of pulses transmitted per second) by \( f_{BB} \). In any of the modes, the PO is heavily duty-cycled which explains the low power consumptions. With a given \( f_{BB} \), the power consumption with DPPM is lower than with OOK despite the fact that the data rates are equal. With baseband clock frequencies of 0.1 and 5 MHz, the EPB with DPPM is 63% and 53% lower compared to OOK, respectively. Fig. 3.10 suggests that, with 48 bits encoded per packet, 64-DPPM outperforms OOK. This suggests that the EPB improvement does not come at the cost of error performance. The smallest power consumption of 67.1 nW is achieved with DPPM and \( f_{BB} = 5 \) MHz.

\(^3\)The data content of a packet does not affect the average PO power consumption much with DPPM – the same amount of pulses is transmitted in every packet regardless of the data. The data, however, affects the operating time of the modulator per packet and its average power consumption therefore.
5.1.4 Uplink measurement results

The uplink capability of the PO-based transmitter was successfully demonstrated over line-of-sight (LOS) uplink ranges of 10 and 30 meters, as reported in [II]. The transmitter was configured to transmit 48-bit DPPM packets with the settings that produced the 67.1-nW power consumption and EPB of 14.0 pJ/bit in packet-mode transmission: \( B = 6 \) bits per symbol, \( N_s = 8 \) symbols per packet and \( f_{BB} = 5 \) MHz. A custom PL-SDD DPPM receiver was used that comprised a laptop, a USRP-2901 software defined radio device [39], an omni-directional \( \lambda/4 \) whip antenna [40] and a 433-MHz SAW filter [41]. In receive mode, a USRP-2901 amplifies, downconverts, filters, digitizes and decimates the received signal and provides it to a host computer [42]. Its noise figure is from 5 to 7 dB [39].

The USRP-2901 outputs digital in-phase (I) and quadrature (Q) signals. An algorithm was written in LabVIEW\(^4\) that demodulates the signal using a PL-SDD scheme similar to that discussed in Section 3.2.4. No baseband clock synchronization nor carrier synchronization was used between the transmitter and the USRP. Without baseband clock synchronization, sampling the DPPM signal once per slot is prone to errors as sampling can occur out of phase. To overcome this issue, 3x oversampling was used. This was performed by setting the USRP to sample the intermediate-frequency (IF) I and Q signals at a sample rate of \( f_s = 3 \cdot f_{BB} = 15 \) MSPS. The local oscillator of the USRP was set to a frequency of \( f_{LO} = 431 \) MHz. The band received by the USRP was thus 423.5–438.5 MHz.

Fig. 5.7 shows parts of the utilized receiver up to a clipper whose purpose is explained shortly. The USRP feeds the I and Q signals to a complex bandpass filter (BPF). This BPF has a bandwidth of 6 MHz and it reduces the signal bandwidth to a band from 431 to 437 MHz. Fig. 5.8(a) shows examples of the received I and Q signals after the complex BPF. To convert the IF signal to baseband, the squarers and the adder calculate the energy of each complex IQ sample. The outputs of the squarers are summed and an energy signal \( E \) is obtained. Signal \( E \) is fed to a moving sum filter (MSF) to perform lowpass filtering. The MSF sums the I and Q signal energy with a window size of three samples which is equivalent to \( T_{BB} \) when \( f_{BB} \) is 5 MHz. Fig. 5.8(b) shows an example of the MSF output, corresponding to the filtered IQ signals of Fig. 5.8(a).

\(^4\)A software by National Instruments.
Notice how the signal of Fig. 5.8(b) resembles the envelope detector output of Fig. 3.9(a) with the exceptions that the signal here is the signal energy instead of amplitude and oversampling is used.

The MSF output is fed to the clipper that takes clips of the MSF output with the length slightly longer than the maximum packet length. The PL-SDD is then performed once per clip. In each clip, the nine highest peaks are interpreted as the centers of the “on” slots.\(^5\) The data content of the received packet is then calculated based on the time delays between the nine peaks. In these uplink measurements, the laptop that controlled the USRP and ran the LabVIEW code was also used to write randomized transmit data to the on-chip memory of the DPPM transmitter chip. The LabVIEW code compared the demodulated data with the intended transmit data to detect packet errors.

A 10-meter link was tested first [II]. In this test setup, interferers occurred occasionally. 40 000 packets were transmitted first and, due to the interfering signals, the PER was only 0.19%. The interferers occurred mostly out of the DPPM signal band but within 423.5–438.5 MHz, the band received by the USRP. Despite the use of the SAW filter, the interferers blocked the intended signal. For better reception, out-of-band blockers could presumably be suppressed using improved bandpass filtering. To emulate the operation without the interferers, the algorithm was changed so that it could detect the interferers from the spectrum of the IF IQ signal. If an interferer was detected, the data was retransmitted. With the packets affected by interferers neglected and retransmitted, 100 000 packets were transferred with zero packet errors. In this measurement, 75 retransmissions were performed due to the presence of an interferer.

Another uplink test was performed with the transmit and receive antennas separated by 30 meters [II]. The setup is shown in Fig. 5.9. 3 000 packets were transmitted over the 30-m range and received correctly except for three packet errors. This suggests that the PER is in the range of 0.1%. The LabVIEW code was configured to save the intended transmit data and IQ signals on a hard drive for error analysis when a packet error occurs. Based on this stored data, it was

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\(^5\)A peak refers to a sample whose preceding and following sample have a lower value.
deduced that the three packet errors were induced by noise, not interferers. The signals of Fig. 5.8 were obtained in the 30-meter test setup. The carrier-to-noise ratio (CNR) was not constant as this was a real environment where, for instance, other radio traffic can impact the reception. Fig. 5.8 depicts the worst observed CNR of approximately 8.2 dB. However, at times, also nearly 10 dB greater values than this were observed.

In Section 2.3, the trade-off between $f_{BB}$ and EPB was discussed. A low EPB can be achieved relatively easily using a high $f_{BB}$ but the cost is a lower SNR and reduced uplink range. Conversely, use of a lower $f_{BB}$ is expected to improve the SNR at the cost of the EPB. This effect is seen in the measurement results of [II]. In [II], the signal reception was tested in the 30-meter uplink measurement setup also with the lower $f_{BB}$ of 100 kHz which resulted in a greater EPB. If the $f_{BB}$ is decreased from 5 MHz to 100 kHz, the noise power in the noise bandwidth is expected to decrease by a factor of 5 MHz/100 kHz = 50, i.e. by $10\cdot\log_{10}(50) \approx 17$ dB. The $f_{BB}$ of the transmitter, generated using an external signal generator, was changed from 5 MHz to 100 kHz. The USRP-based receiver was adjusted accordingly: the bandwidth of the complex BPF was decreased by a factor of 50 from 6 MHz to 120 kHz and the window size of the MSF was increased from 3 to 150 samples to match the fiftyfold $T_{BB}$. Data transmission was performed and the I and Q signals received in this new configuration were captured from the USRP. Approximate CNR was then calculated. The observed CNR values increased by 16 dB compared to the case with $f_{BB} = 5$ MHz which corresponds quite well with the expected decrease in noise power [II]. As stated in [II], it can be estimated by path loss calculations that this amount of improvement in CNR would increase the line-of-sight uplink range from 30 to approximately 200 meters.

This test in [II] showed the trade-off between the EPB and SNR. As shown in Table 5.1, using 64-DPPM with an $f_{BB}$ of 100 kHz consumes 517 pJ/bit while only 14 pJ/bit is consumed with an $f_{BB}$ of 5 MHz. However, the higher EPB
enabled a higher CNR in the previous measurement, described in [II]. A higher CNR can be worth extra consumed energy whenever a higher uplink range or a lower error probability is required. The EPB is not a good metric for energy efficiency because it completely neglects the SNR and the effect of bandwidth which was briefly observed here. This trade-off should be accounted for in an energy efficiency FOM. It justifies the inclusion of the noise bandwidth in the FOM discussed in Section 2.4 and [I].

5.1.5 FOM calculations

The following FOM calculations have not been included in [II] or the other publications included in this thesis. However, the resulting FOM and link strength with the 64-DPPM mode were used in the transmitter comparison in [I] which was published after [II]. The FOM with OOK is calculated using (2.25). We consider the performance achieved with the continuous-mode transmission with the baseband clock frequency of $f_{BB} = 5$ MHz. Hereby, the noise bandwidth is $BW_N = 5$ MHz. In this mode, the TX current consumption was $119.3 \mu A$ from a 1.2-V supply for which the power consumption is $P_{TX} = 1.2 \text{ V} \cdot 119.3 \mu A \approx 143.2 \mu W$. OOK requires a $\gamma_{req}$ of $16.1$ dB $\approx 40.7$ for the targeted BER of $10^{-5}$, or the corresponding PER of $4.8 \cdot 10^{-4}$ with 48-bit packets (see Fig. 3.13). The peak output power is $-25$ dBm which is approximately $3.16 \mu W$ and we consider noise temperature $T = 298$ K. Substituting these values to (2.25) yields

$$FOM_{OOK} = \frac{143.2 \mu W \cdot k \cdot 298 \text{ K} \cdot 5 \text{ MHz} \cdot 40.7}{5 \text{ Mbps} \cdot 3.16 \mu W} \approx 7.59 \cdot 10^{-18} \text{ J/bit} \approx -171.2 \text{ dBJ/bit.}$$

With 64-DPPM, we use (2.21) to calculate the FOM.\textsuperscript{6} The DPPM packet error ratios in Fig. 3.13 assume the use of PL-SDD at a receiver and, consequently, packet-mode transmission. Thus, we consider the performance of the transmitter when transmitting 64-DPPM packets. According to Fig. 3.13, $\gamma_{req}$ is $14.8$ dB. The EPB was $14.0$ pJ/bit according to the measurements where 64-DPPM data was transmitted in packet-mode. We consider the achieved performance with $f_{BB} = 5$ MHz for which the noise bandwidth is $5$ MHz. The maximum SNR achievable with the output signal is obtained with (2.24) which, with the peak output power of $-25$ dBm $\approx 3.16 \mu W$, yields

$$\gamma_{max} = \frac{3.16 \mu W}{k \cdot 298 \text{ K} \cdot 5 \text{ MHz}} \approx 1.54 \cdot 10^8 \approx 81.9 \text{ dB}. \quad (5.1)$$

Substituting the EPB, $\gamma_{req}$ and $\gamma_{max}$ to (2.21) yields

$$FOM_{64-DPPM} = 10 \cdot \log_{10} \left( \frac{14.0 \text{ pJ/bit}}{1 \text{ J/bit}} + 14.8 \text{ dB} - 81.9 \text{ dB} \right) \approx -175.6 \text{ dBJ/bit.}$$

\textsuperscript{6}All the FOM equations are expected to yield the same result when proper values are used in the calculation. The use of (2.21) does not bring any unfair advantage here.
Thus, the FOMs with OOK and 64-DPPM are –171.2 and –175.6 dBJ/bit, respectively. The FOM with 64-DPPM is 4.4 dB better compared to OOK. This improvement is explained by two features. Firstly, the transmitter consumes approximately 3 dB less energy per transmitted bit in the 64-DPPM mode. Secondly, an additional improvement of 1.3 dB comes from the $\gamma$ requirement – 64-DPPM requires less carrier power than OOK for the targeted PER (see Fig. 3.13) or, conversely, achieves lower PER with an equal $\gamma$. With (2.19), it can be calculated that the link strengths with OOK and 64-DPPM are 65.8 and 67.1 dB, respectively. As the peak output powers and noise bandwidths are equal with both modulation schemes, the transmitter is expected to achieve a slightly improved uplink performance with 64-DPPM despite the lower EPB.

5.2 Direct-modulation narrowband DPPM transmitter

5.2.1 Overview

The second narrowband transmitter prototype, presented in [I], is based on the direct-modulation architecture of Fig. 2.7(a). It supports OOK and DPPM and includes pulse-shaping capability for reducing the PSD of the signal sidelobes in the spectrum and, consequently, the occupied bandwidth (OBW). Fig. 5.10 depicts the OOK and DPPM baseband signals with 10x oversampled pulse shaping and Fig. 5.11 shows the block diagram of the system [I]. A carrier frequency of 434 MHz is used. A 2-axis gesture sensor interface is included on the same chip that provides two digital data streams at a sampling rate of roughly $7_{47.2}$ Hz. The sensor is discussed in Section 6.2. An SPI and a memory register are included for controlling the chip and providing trim bits for the circuits. A 4.4-MHz ring oscillator acts as the clock reference. The baseband clock frequency is $f_{BB} = 440$ kHz and the RO frequency is tenfold as 10x oversampled pulse shaping is used. The carrier frequency is generated with the RF oscillator and a 50-Ω load is driven by the power amplifier. The transmitter supports

$\gamma$ The sampling rate is not defined by a high-precision frequency generator but by an on-chip ring oscillator. Some frequency variation can thus be expected.

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OOK and DPPM. In the case of both modulation schemes, the modulator block, denoted as OOK/DPPM MOD, generates control signals for the RF oscillator and PA. The control signals are used for duty-cycling and RF output amplitude control.

5.2.2 RF oscillator and power amplifier

Fig. 5.12 shows the schematic of the 434-MHz Pierce RF oscillator and the PA. The staggered enable signals (EN1–EN5) are depicted at the bottom right. When enabled and disabled, the oscillator and PA both switch on and off rapidly. This enables efficient duty-cycling and, hence, energy-efficient transmission of DPPM and OOK signals.

The Pierce oscillator contains two oscillator cores, each consisting of an inverter and header transistors $M_{OSC0}$–$M_{OSC5}$. The use of two cores enables sufficient driving strength and fast start-up in slow corners. If a slower start-up is acceptable, one of the cores may be disabled to conserve power. Active power
ULP transmitter prototypes in 180 nm CMOS
consumption can furthermore be reduced by leaving part of the header transis-
tors disabled. The enable and disable signals are dynamic and the oscillator is
switched off during transmission of zeros and “off” slots with OOK and DPPM,
respectively. The supply voltage of the oscillator is $V_{DD_{OSC}} = 0.9$ V.

The carrier frequency is defined by tunable on-chip capacitors and an off-
chip inductor. The tunable capacitors are identical to those used with the
PO-based transmitter prototype, described in Section 5.1.1. $C_{coarse}$ and $C_{med}$
are controlled with 3 and 4 tuning bits, respectively. Two fine tuning capacitors
($C_{fine}$) are used. In a similar fashion to the PO-based transmitter, each is
connected between the node VP or VN and ground for smaller frequency tuning
steps [38]. The fine tuning capacitors have individual tuning bits, 6 bits per
capacitor. To produce the carrier frequency of 434 MHz, an external 27-nH
inductor is used in the LC tank.

The PA consists of the NMOS transistor $M_{PA}$, cascode transistors $M_{CAS0}–
M_{CAS6}$, choke inductor $L_1$, DC block $C_1$ and a matching network consisting of
$C_M$ and $L_M$. The cascode transistors are used to control the drain current $I_D$
and the RF output amplitude. The gate voltages of these transistors (nodes
denoted as ENA_P<6:0>) are dynamically controlled with the staggered en-
able signals which enables pulse shaping. The more cascode transistors are
active, the greater the RF output amplitude is. Each of the seven gate voltages
ENA_P<6:0> can be multiplexed from any of the five enable signals EN<1–EN<5
or from logic zero. This enables pulse shaping with various envelope shapes.
The supply voltage of the PA is $V_{DD_{PA}} = 0.9$ V. The enable signals are obtained
from the modulator whose supply voltage is 1.2 V and, thus, their voltage levels
are 0 and 1.2 V.

5.2.3 Modulator and 4.4-MHz ring oscillator

The modulator block of the second prototype possesses the same functionality as
the DPPM modulator of the first prototype, described in Section 5.1.2. However,
some modifications have been made for generating the multiple control signals
used in the pulse shaping. The modified modulator is shown in Fig. 5.13(a). The DPPM modulator, shown on the left, is equivalent to that of Section 5.1.2.
It generates the DPPM baseband signal where a pulse the length of $T_{BB}$ is
generated each time an “on” slot is generated. The DPPM packet format is
identical to the first transmitter prototype. A packet begins with a start pulse,
guard slots are utilized and one pulse is transmitted per symbol after the start
pulse. The value of $B$ can be set from 1 to 6 using SYM values limited from 0 to
$2^B–1$. Thus, DPPM schemes from 2-DPPM to 64-DPPM are supported.

The DPPM modulator is followed by a state machine that generates the
staggered enable signals, EN<1–EN<5>. Fig. 5.13(b) depicts the operation of the
state machine. The state is toggled with the 10x clock frequency, $10 \cdot f_{BB}$. The
values of outputs EN<1–EN<5> are defined by the different states. If BIT' is low, the
state is such that the outputs are low. If BIT' is high, the state is incremented
Figure 5.13. (a) Functional block diagram of the modulator block of the second transmitter prototype, and (b) generation of the staggered enable signals, EN<sub>1</sub>–EN<sub>5</sub>, using a state machine. Subfigure (a) derived from [III], used under CC BY 4.0 / removed: PO-based transmitter; added: two muxes, state machine and signal mux with related signals.

Clock cycle by clock cycle and the outputs EN<sub>1</sub>–EN<sub>5</sub> are first sequentially enabled starting from EN<sub>1</sub> and then disabled starting from EN<sub>5</sub>. The signal NEXTBIT<sup>′</sup> is used in OOK mode. In DPPM mode, it is always low. In OOK mode, if BIT<sup>′</sup> and NEXTBIT<sup>′</sup> are both high, the state is changed to a hold state when EN<sub>1</sub>–EN<sub>5</sub> have all been set to high. In Fig. 5.13(b), this occurs at the time instant <i>t</i><sub>1</sub>. In the hold state, EN<sub>1</sub>–EN<sub>5</sub> are all maintained high when multiple ones are transmitted in sequence. This maintains the RF output amplitude high as depicted in Fig. 5.10. The signal multiplexer after the state machine multiplexes the enable signals to the outputs ENA_OSC, DISA_OSC1<sub>&lt;5:0&gt;</sub>, DISA_OSC2<sub>&lt;5:0&gt;</sub> and ENA_PA<sub>&lt;6:0&gt;</sub> that control the RF oscillator and PA.

Fig. 5.14 shows the current-starved ring oscillator that acts as the 4.4-MHz clock reference. Its core is a three-stage ring oscillator. Two buffer stages are used. The frequency can be tuned by adjusting the bias current $I_B$, provided by the digitally controlled bias current generator (BCG) of Fig. 5.11. To disable the RO when the transmitter is off, e.g. between transmission of packets, $I_B$ can be cut off by setting the ENA_BB input of the BCG low. For this prototype, no dynamic on-chip control of the ENA_BB signal was implemented but, for RO duty-cycling in measurements, it can be toggled dynamically using the SPI. The supply voltage of the ring oscillator is 1.2 V.
5.2.4 Measurement results

The second transmitter prototype was implemented in a 180 nm CMOS process. Fig. 5.15 shows a micrograph of the implemented blocks with the 4.4-MHz RO with its BCG, modulator (MOD) and transmitter front-end (TXFE), consisting of the RF oscillator and PA, on the right. The blocks near the left edge are related to the gesture sensor that is presented later in Section 6.2.

The key measurement results have been presented in [I]. An external inductor of 27 nH was used in the LC tank of the RF oscillator. With the carrier frequency tuned to 434 MHz, the RF oscillator and PA consume 316 µW and 1337 µW of power, respectively, when the carrier is generated continuously. The standby powers of the blocks are 0.1 nW and 0.9 nW, respectively. The maximum output power delivered to a 50-Ω load is –2.1 dBm when all the PA cascode transistors are active. Thus, the peak global efficiency and peak PA drain efficiency are 37% and 46%, respectively. Fig. 5.16 shows a graph not included in [I]: the frequency of the RF oscillator against the frequency tuning word TUNE_FREQ. The frequency tuning range is from 484.6 to 426.5 MHz. The descending frequency steps are also shown and their magnitudes are lower than 41 kHz throughout the range.

Fig. 5.17 shows the effect of the pulse shaping on the RF output in the time domain. The sub-plots (a) and (b) show the output waveform without and with
Figure 5.16. Frequency of the Pierce RF oscillator versus the frequency tuning code. Descending frequency step sizes are shown in grey.

Figure 5.17. Output RF waveform of the transmitter (a) without and (b) with pulse shaping showing OOK data pattern 1011. The RF waveform with DPPM during an “on” slot is the same as a sole bit 1 with OOK. Obtained from [I], used under CC BY 4.0.

shaping, respectively. Fig. 5.18 shows the effect in the frequency domain. In the spectrum measurements, a repeating OOK data pattern of 1010 was used which produces the widest occupied bandwidth. This RF waveform is also equivalent to that produced by DPPM when guard slots are utilized and the data stream consists of zero-only symbols.

ETSI allows an occupied bandwidth of 1.75 MHz from 433.04 to 434.79 MHz [43]. The occupied bandwidth was measured using a Keysight N9041B with the span set to 32 MHz to include the major sidebands. The frequency of the 4.4-MHz RO was set to 4.398 MHz. Thus, the baseband clock frequency was $f_{BB} = 439.8$ kHz. With this $f_{BB}$ and the 1010 data pattern, the measured OBW is 1.65 MHz with the pulse shaping, i.e. lower than the 1.75-MHz requirement.

Table 5.2 shows the measurement results with continuous-mode transmission of data using OOK, 16-DPPM and 64-DPPM. The power consumption includes the power consumed by the RF oscillator, PA, 4.4-MHz RO and the modulator. To measure the average consumption with OOK, equiprobable bits were used, i.e. 50% of the data bits were ones. With DPPM, the power consumption is significantly lower compared to OOK but also the data rate is reduced. The
ULP transmitter prototypes in 180 nm CMOS

**Figure 5.18.** Output spectrum of the transmitter with and without pulse shaping when OOK data pattern 1010 is transmitted continuously. VBW is the video bandwidth. Obtained from [I], used under CC BY 4.0.

**Table 5.2.** Measurement results with the second transmitter prototype when modulated data is transmitted continuously with pulse shaping

<table>
<thead>
<tr>
<th>Modulation</th>
<th>OOK</th>
<th>16-DPPM</th>
<th>64-DPPM</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bits/symbol</td>
<td>1</td>
<td>4</td>
<td>6</td>
</tr>
<tr>
<td>Power Consumption (µW)</td>
<td>736</td>
<td>140</td>
<td>41.1</td>
</tr>
<tr>
<td>Average Data Rate (kbps)</td>
<td>440</td>
<td>185 †</td>
<td>78.8 †</td>
</tr>
<tr>
<td>Average Output Power (dBm)</td>
<td>-6.2</td>
<td>-14.3</td>
<td>-19.8</td>
</tr>
<tr>
<td>TX Efficiency</td>
<td>32.6%</td>
<td>26.5%</td>
<td>25.5%</td>
</tr>
<tr>
<td>EPB (nJ/bit)</td>
<td>1.67</td>
<td>0.76</td>
<td>0.52</td>
</tr>
</tbody>
</table>

Table obtained from [I], used under CC BY 4.0.

† With DPPM with guard slot, data rate is $2 \cdot f_{BB} \cdot B/(2^B + 3)$ as per (3.12).

average output power is lower with DPPM but it can be noted again that this does not imply inferior uplink performance with DPPM as was discussed in Section 5.1.3. DPPM achieves a significantly lower EPB than OOK. With OOK, the global efficiency is slightly higher. This is because the TX efficiency is higher with higher output power and, with OOK, the output power is maintained at maximum for longer periods of time when several ones are transmitted in sequence, as can be seen in Fig. 5.17(b) on the right. The lower efficiency with DPPM is acceptable because of its other benefits: the PER is expected to be lower (see Fig. 3.10) despite the lower EPB.

In [I], the power and energy consumption of the TX were also reported with data transmitted in packet-mode using 64-DPPM. In this measurement, the packet size was set to 48 bits with $N_s = 8$ symbols transmitted per packet and $B = 6$ bits encoded per symbol. In addition to the eight data symbols, a timing reference symbol was transmitted as a part of the packet. This counteracts timing uncertainty related to the on-chip 4.4-MHz RO. This symbol has a fixed duration, in this case $17 \cdot T_{BB}$, and it carries information about $T_{BB}$ and $f_{BB}$ to a DPPM receiver. The receiver can use this timing information for decoding the
data [IV]. With the start pulse, each packet hence contained a total of 10 pulses. Guard slots were utilized. A packet size of 48 bits is enough for carrying the two raw 24-bit gesture sensor output samples. As the sensor accuracy is lower than 24 bits, the data could also be truncated and a part of the 48 bits could be used for error detection or correction. The sensor sample rate is roughly 47.2 SPS. Hence, 47.2 packets were transmitted per second.

As explained in [I], to implement the packet-mode transmission, an FPGA was used for controlling the transmitter through the SPI. Fig. 5.19 shows some relevant signals from the measurement. Transmission of one packet is depicted. The top subfigure shows the RF output, SCK line (i.e. the clock line of the SPI), and the 4.4-MHz RO output. Three SPI bursts were used for basic control. The first one switches on the 4.4-MHz RO at the time instant \( t_0 \) by enabling the BCG. The second one triggers the transmission of a packet at \( t_1 \). The third one switches off the 4.4-MHz RO at \( t_2 \) to conserve power until the next packet is transmitted. In the figure, the data in all the symbols is 111111\(_2\) which results in the longest symbol duration. The figure thus depicts the maximum packet duration which is roughly 1.2 ms from the first RF pulse to the last. The bottom subfigure shows the period of the RO. It settles well before the transmission is triggered at the time instant \( t_1 \). If a shorter transmission time was desirable, a lower \( B \) could be used with the number of symbols \( N_s \) adjusted accordingly, for instance, \( B = 4 \) and \( N_s = 12 \) or \( B = 2 \) and \( N_s = 24 \).

In this packet-mode measurement with 47.2 packets transmitted per second, the average power consumption of the transmitter was 1.56 \( \mu \)W. The shares of the RF oscillator, PA, modulator and 4.4-MHz RO are 346 nW, 1.05 \( \mu \)W, 46 nW and 116 nW, respectively. The modulator and RO consume considerably less

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8 The RO frequency can vary due to e.g. supply voltage and temperature variation which impacts the baseband clock period \( T_{BB} \) and the DPPM pulse timing. If the variation is excessive and a receiver does not know the frequency, reception of the DPPM packets may be impossible. The extra symbol delivers the frequency information to the receiver.
power than the RF oscillator and PA. The data rate is \(47.2 \cdot 48 \text{ bps} = 2265.6 \text{ bps}\). Additional energy is required to transmit the start pulse and the timing reference symbol. For this, the EPB is slightly greater compared to continuous-mode transmission, 0.69 nJ/bit. For a minor EPB reduction, the modulator and RO power consumption could be roughly halved by improving the RO duty-cycling by modifying the control code programmed to the FPGA. In this test, the RO duty cycle was not made data-dependent. Instead, the RO was on for 1.4 ms per packet despite the data content. While Fig. 5.19 shows the longest packet, the packet duration is roughly half of this on average with random data. For this, the RO on-time would only need to be approximately 0.7 ms on average.

5.2.5 FOM calculations

In [I], the FOM in the 64-DPPM mode was considered but the FOM was not estimated for the OOK mode. These FOMs are obtained as follows. The FOM in the 64-DPPM mode is calculated in a similar fashion to that of the first transmitter prototype in Section 5.1.5. The performance in the packet-mode transmission is considered as the PER in (3.23), evaluated in Fig. 3.10 and Fig. 3.13, assumes the use of PL-SDD at a receiver. The EPB was 0.69 nJ/bit in the measurement where 64-DPPM data was transmitted in 48-bit packets. It is to be noted that, in the measurements with this transmitter, the timing reference symbol was added to the packet format which affects the error probability. The addition of the timing reference increases the numbers of the “on” and “off” slots by one and 16, respectively. Thus, we have \(N_O = 10\) and \(N_Z = 528\). However, recalculating the PER in (3.23) with the new values of \(N_O\) and \(N_Z\) shows that \(\gamma_{req}\) only increases by approx. 0.05 dB, and that it is 14.8 dB also with the timing reference.

In addition to the slightly changed packet format, pulse shaping is used which could affect the PER. To evaluate its effect, waveform-level PER simulations were performed according to Section 3.3. The emulated 64-DPPM RF signal was now generated using the pulse envelope of Fig. 5.17(b) instead of a flat envelope. Correspondingly, this shaped envelope was also applied to the FIR filter coefficients of the BPF in Fig. 3.12(a) which had a flat envelope in the simulations of Section 3.3. With this filtering, the \(\gamma\) required for \(\text{PER} = 4.8 \cdot 10^{-4}\) was still 14.8 dB. Moreover, the \(\gamma\) in the sampled envelope detector output was still determined as \(\gamma = E_{\text{sig}}/E_n\) where \(E_{\text{sig}}\) is now the energy of a shaped RF waveform representing an “on” slot and \(E_n\) the energy of the noise in bandwidth \(f_{BB}\) over the time interval \(T_{BB}\). This suggests that the noise bandwidth is not affected by the shaping. For the above reasons, despite the slightly modified packet format and the use of the shaping, we use \(\gamma_{req} = 14.8 \text{ dB}\) and a noise bandwidth of \(f_{BB}\), i.e. 440 kHz, in the FOM calculation.

Because pulse shaping is used, the peak output power of \(-2.1 \text{ dBm}\) cannot be directly used in the FOM calculation. Instead, we use the effective output power during a DPPM “on” slot whose value, \(P_{\text{out,eff}} = -4.5 \text{ dBm} \approx 354.8 \mu\text{W},\)
was obtained in [I]. With this, the maximum SNR achievable with the output signal is obtained with (2.24) as

\[ \gamma_{\text{max}} = \frac{354.8 \, \mu W}{k \cdot 298 \, K \cdot 440 \, \text{kHz}} \approx 1.959 \cdot 10^{11} \approx 112.9 \, \text{dB}. \]

Substituting EPB = 0.69 nJ/bit, \( \gamma_{\text{req}} = 14.8 \, \text{dB} \) and \( \gamma_{\text{max}} = 112.9 \, \text{dB} \) to (2.21) yields the FOM for the 64-DPPM mode, reported in [I] and calculated here as

\[
FOM_{\text{64-DPPM}} = 10 \cdot \log_{10} \left( \frac{0.69 \, \text{nJ/bit}}{1 \, \text{J/bit}} \right) + 14.8 \, \text{dB} - 112.9 \, \text{dB} \approx -189.7 \, \text{dBJ/bit}.
\]

The FOM was compared in [I] with state-of-the-art ULP transmitters. The comparison results are also discussed in this thesis in Section 5.3.

Regarding OOK, there is a challenge related to the pulse shaping implementation: the symbol energies with OOK related to bit 1 are not equal. The preceding and following bit impact the envelope of the waveform (see Fig. 5.17(b)). As the preceding and following bit can each assume one of two values, bit 1 may be represented by one of four different waveforms. This affects the envelope detector output distribution at a receiver (see Fig. 3.12(a)) and, furthermore, the BER and PER. In this work, DPPM is the main modulation scheme to be used with the transmitter. For that, we shall not analyze what is the best achievable error performance with an OOK signal shaped in this manner. Instead, we simply use the SNR requirement of OOK, \( \gamma_{\text{avg}, \text{req}} = 13.1 \, \text{dB} \approx 20.4 \), in the calculation to get a rough estimate of the performance in the OOK mode. With the pulse shaping, the consumed power was 736 \( \mu \text{W} \) at the data rate of 440 kbps. The average output power was \( P_{\text{out}} = -6.2 \, \text{dBm} \approx 239.9 \, \mu \text{W} \). Substituting these values to (2.26) yields

\[
FOM_{\text{OOK}} = \frac{736 \, \mu \text{W} \cdot k \cdot 298 \, K \cdot 440 \, \text{kHz} \cdot 20.4}{440 \, \text{kbps} \cdot 239.9 \mu \text{W}} \approx 2.58 \cdot 10^{-19} \, \text{J/bit} \approx -185.9 \, \text{dBJ/bit}.
\]

Thus, the FOMs with 64-DPPM and OOK are -189.7 and approximately -185.9 dBJ/bit, respectively. In this case, the 64-DPPM mode outperforms the OOK mode by approximately 3.8 dB in terms of energy efficiency. As stated in Section 5.1.5, the reasons are the lower EPB with DPPM and the lower \( \gamma_{\text{req}} \) when a DPPM receiver utilizes the PL-SDD.

In the 64-DPPM mode, the link strength is obtained with (2.27). Using \( P_{\text{out,eff}} \) in the calculation, we get the result reported in [I], obtained as

\[
LS_{\text{64-DPPM}} = \frac{354.8 \, \mu \text{W}}{k \cdot 298 \, K \cdot 440 \, \text{kHz} \cdot 10^{14.8/10}} \approx 6.49 \cdot 10^9 \approx 98.1 \, \text{dB}.
\]

---

9 The output power was -7.5 dBm when a 1010 pulse train was transmitted with pulse shaping. As a pulse was included in every other slot, \( P_{\text{out,eff}} \) is twofold, i.e. 3 dB greater.

10 The preceding bit determines if the waveform starts with a rising envelope or with the maximum amplitude. The following bit determines if the waveform ends with a falling envelope or remains at the maximum amplitude.
5.3 Energy efficiency comparison of state-of-the-art ULP narrowband transmitters

Table 5.3, adapted from [I], shows an energy efficiency comparison between the narrowband transmitters implemented in this thesis, i.e. [I] and [II], and other ULP narrowband transmitters. The table includes the state-of-the-art transmitters which were reviewed in Section 2.5.4. Furthermore, it includes additional OOK transmitters from [15], [31], [33] and [44], a BPSK transmitter from [45], BFSK transmitters from [30] and [46], and Bluetooth Low Energy (BLE) transmitters from [25], [47] and [48]. Such transmitter properties are listed that can be used for the FOM calculation using (2.26).\footnote{Note that the FOM of a packet-mode DPPM transmitter cannot be calculated using (2.26) unless one derives the required average-signal-power-to-noise-power ratio $\gamma_{avg,req}$. With packet-mode DPPM, it is more straightforward to calculate the EPB and $\gamma_{max}$ separately and calculate the FOM using (2.21) as was done in Sections 5.1.5 and 5.2.5.} $P_{TX}$ and $P_{out}$ are the average power consumption and output power, respectively, when modulated data is transmitted at the data rate $R_b$. The noise bandwidth is equal to $R_b$ with BPSK, OOK and BFSK. With these modulation schemes, $\gamma_{avg,req}$ is 9.6, 13.1 and 13.4 dB, respectively, for the targeted error performance, i.e. $BER = 10^{-5}$ and the corresponding PER of 4.8·$10^{-4}$ with 48-bit packets. BLE transmitters use Gaussian frequency-shift keying (GFSK) with a modulation index $h$ between 0.45 and 0.55 and a bandwidth-bit period product $BT$ of 0.5 [49]. Assuming noncoherent reception and the use of $h = 0.5$, the BER with GFSK is the same as with BFSK [50]. Thus, with the BLE transmitters, $\gamma_{avg,req}$ of 13.4 dB is assumed.

The noise bandwidth with GFSK is obtained as $BW_N = \sqrt{\pi/(4\cdot\log_{10}(2))\cdot BT\cdot R_b}$ [51], i.e. as $BW_N \approx 0.8076\cdot R_b$. The table also lists the power efficiencies that have been calculated as $\eta = \frac{P_{out}}{P_{TX}}$. The link strengths can be calculated with (2.28) which, using decibel values, can also be written as

$$LS = (P_{out})_{dBm} - 10 \cdot \log_{10} \frac{k \cdot T \cdot BW_N}{1 \text{ mW}} - (\gamma_{avg,req})_{dB}. \quad (5.2)$$

The works in Table 5.3 are listed in the order of the achieved energy efficiency FOMs from the best to worst. The most energy-efficient transmitters are at the top of the list. The 64-DPPM transmitter [I] of this thesis achieves the best energy efficiency FOM, enabled by power-efficient circuitry and the use of an energy-efficient modulation scheme. Desirable FOMs are also achieved by BPSK and high-output-power BLE transmitters. The transmitters at the bottom of the list generally suffer from low power efficiency, use of a modulation scheme that is not very energy-efficient, or both. It can be seen in the table that the energy efficiency FOM does not correlate with the power consumption or EPB. It is rather the power efficiency and the choice of modulation scheme that matter. Although a low power consumption and EPB are desirable, they are not very comprehensive metrics of the overall transmitter performance or energy efficiency. Many of the transmitters in the table with the lowest EPBs also have very low power efficiencies which implies that they consume excessive
The new energy efficiency FOM which was derived in Section 2.4.
4) Average duty cycle is 50% with OOK and 3.0% with 64-DPPM with guard slots.
5) EPBs of [I] and [II] are 0.69 nJ/bit and 14 pJ/bit, respectively, when packet-mode transmission is used. The link strengths of 98.1 and 67.1 dB here apply to packet-mode transmission. See Sections 5.2.5 and 5.1.5 for the EPB, LS and FOM calculations.
6) FOM of DPPM TX evaluated based on performance in packet-mode transmission, not continuous-mode.
7) PO-based TX. Output power as limited by fixed coil-antenna.

An observation can be made in the results of the ultra-low power BLE trans-

power with respect to the achieved link strength. Also their link strengths are relatively low. Particularly the PO-based transmitters in [II], [28] and [48] appear to have had challenges with power efficiency. A part of the problem with these PO-based works is that, in a PO, a loop antenna is utilized as the inductor in the LC tank. The antennas in these works are electrically small compared to the carrier frequencies. The radiation efficiency [52] of an electrically small antenna is low, i.e. it only radiates a minor portion of the power that is delivered to it. This results in a low antenna gain, loss of power and, consequently, a low power efficiency.

An observation can be made in the results of the ultra-low power BLE trans-

Table 5.3. Energy efficiency comparison of ULP narrowband transmitters

<table>
<thead>
<tr>
<th>Modulation</th>
<th>$f_c$ [MHz]</th>
<th>$P_{TX}$ [µW]</th>
<th>$P_{out}$ [dBm]</th>
<th>$R_s$ [kbps]</th>
<th>$\eta$ [%]</th>
<th>LS [dB]</th>
<th>FOM [dBJ/bit]</th>
</tr>
</thead>
<tbody>
<tr>
<td>[I] 64-DPPM$^4$</td>
<td>434</td>
<td>41</td>
<td>-19.8</td>
<td>78.8</td>
<td>522$^5$</td>
<td>25.5</td>
<td>98.1$^6$</td>
</tr>
<tr>
<td>BPSK</td>
<td>400</td>
<td>67</td>
<td>-17.5</td>
<td>1000</td>
<td>67</td>
<td>26.5</td>
<td>86.8</td>
</tr>
<tr>
<td>GFSK/BLE</td>
<td>2400</td>
<td>1550</td>
<td>-3.3</td>
<td>1000</td>
<td>1550</td>
<td>30.2</td>
<td>98.1</td>
</tr>
<tr>
<td>[12] BPSK</td>
<td>2450</td>
<td>530</td>
<td>-11</td>
<td>1000</td>
<td>530</td>
<td>15.0</td>
<td>93.2</td>
</tr>
<tr>
<td>OOK$^4$</td>
<td>440</td>
<td>0</td>
<td>20000</td>
<td>16.5</td>
<td>9.6</td>
<td>76.3</td>
<td>-184.1</td>
</tr>
<tr>
<td>BPSK</td>
<td>400</td>
<td>330</td>
<td>-15</td>
<td>20000</td>
<td>16.5</td>
<td>9.6</td>
<td>76.3</td>
</tr>
<tr>
<td>[11] BFSK</td>
<td>401</td>
<td>90</td>
<td>-17</td>
<td>200</td>
<td>450</td>
<td>22.2</td>
<td>90.4</td>
</tr>
<tr>
<td>[44] OOK$^4$</td>
<td>440</td>
<td>2600</td>
<td>-3.2</td>
<td>40000</td>
<td>16.5</td>
<td>9.6</td>
<td>76.3</td>
</tr>
<tr>
<td>BPSK</td>
<td>920</td>
<td>700</td>
<td>-10</td>
<td>5000</td>
<td>140</td>
<td>14.3</td>
<td>83.2</td>
</tr>
<tr>
<td>OOK$^4$</td>
<td>433</td>
<td>160</td>
<td>-17</td>
<td>1000</td>
<td>160</td>
<td>12.5</td>
<td>83.7</td>
</tr>
<tr>
<td>[33] OOK$^4$</td>
<td>433</td>
<td>518</td>
<td>-12.7</td>
<td>10000</td>
<td>52</td>
<td>10.4</td>
<td>78.1</td>
</tr>
<tr>
<td>[26] OOK$^4$</td>
<td>401</td>
<td>71</td>
<td>-24</td>
<td>200</td>
<td>360</td>
<td>5.6</td>
<td>83.8</td>
</tr>
<tr>
<td>[37] BPSK</td>
<td>480</td>
<td>170</td>
<td>-20</td>
<td>1000</td>
<td>170</td>
<td>5.9</td>
<td>80.5</td>
</tr>
<tr>
<td>[31] BFSK</td>
<td>2450</td>
<td>600</td>
<td>-17</td>
<td>20000</td>
<td>30</td>
<td>3.3</td>
<td>70.7</td>
</tr>
<tr>
<td>[II] 64-DPPM$^4$</td>
<td>434</td>
<td>10.4</td>
<td>-40$^7$</td>
<td>895.5</td>
<td>11.6$^6$</td>
<td>1.0</td>
<td>67.1$^5$</td>
</tr>
<tr>
<td>OOK$^4$</td>
<td>432</td>
<td>248</td>
<td>-24</td>
<td>1000</td>
<td>248</td>
<td>1.6</td>
<td>76.7</td>
</tr>
<tr>
<td>GFSK/BLE</td>
<td>2400</td>
<td>606</td>
<td>-23$^4$</td>
<td>1000</td>
<td>606</td>
<td>0.8</td>
<td>78.0</td>
</tr>
<tr>
<td>[28] OOK$^4$</td>
<td>2400</td>
<td>191</td>
<td>-29$^7$</td>
<td>5000</td>
<td>38</td>
<td>0.7</td>
<td>64.7</td>
</tr>
<tr>
<td>BFSK</td>
<td>374</td>
<td>500</td>
<td>-26$^5$</td>
<td>5000</td>
<td>75</td>
<td>0.7</td>
<td>67.5</td>
</tr>
</tbody>
</table>

Table adapted from [I], used under CC BY 4.0 / added more publications and columns.
1) The power consumption of some circuit blocks depends on the carrier frequency. The use of a high carrier frequency may degrade energy efficiency.
2) Average during continuous transmission of modulated data.
3) The new energy efficiency FOM which was derived in Section 2.4.
4) Average duty cycle is 50% with OOK and 3.0% with 64-DPPM with guard slots.
5) EPBs of [I] and [II] are 0.69 nJ/bit and 14 pJ/bit, respectively, when packet-mode transmission is used. The link strengths of 98.1 and 67.1 dB here apply to packet-mode transmission. See Sections 5.2.5 and 5.1.5 for the EPB, LS and FOM calculations.
6) FOM of DPPM TX evaluated based on performance in packet-mode transmission, not continuous-mode.
7) PO-based TX. Output power as limited by fixed coil-antenna.
miters: undesirable FOMs have been achieved by the BLE transmitter of [25] in the low-output-power mode and by the BLE transmitter of [48] whose output power is low. This could be related to the fact that a continuously running PLL is generally used to generate a GFSK signal with the smooth frequency transitions. The use of GFSK and a continuously running power-consuming PLL might not be the optimum choice in terms of energy efficiency in low-output-power scenarios. Energy could be saved by setting the transmit frequency using a duty-cycled PLL instead. However, that may generally prevent the use of GFSK. It is to be noted that, even if the use of GFSK causes a higher power consumption, it has a significant benefit that could be worth the extra power – a GFSK signal causes significantly less interference outside the signal band compared to non-shaped OOK, BPSK and BFSK signals [6]. On the other hand, the use of OOK, BPSK, BFSK, PPM or DPPM with pulse-shaping could offer decent spectral quality compared to non-shaped signals while enabling lower LO power consumption compared to GFSK. The transmitters in [I] and [12] utilize pulse-shaping and achieve decent energy efficiencies.

Fig. 5.20, adapted from [I], shows the link strengths achieved by the works included in Table 5.3 versus the EPBs. It was stated in [I] that the main energy-related challenge in radio transmitter design is in achieving a given link strength with the minimum energy consumed per bit. Correspondingly, a general trend is visible in the figure: the link strengths are lower with the lower EPBs. It can be seen that a high EPB is not necessarily a sign of low energy efficiency. Instead, it may mean that the link strength is higher. For this, the EPB is not a proper metric for energy efficiency – it does not properly consider what is achieved with the consumed energy. The thick green diagonal line denotes the state of the art and corresponds to an FOM of –190 dBJ/bit. The higher the link strength is compared to the EPB, the better the FOM of a transmitter is, and the closer the transmitter is to this line. A high and undesirable FOM thus means that a transmitter consumes high energy per bit compared to the uplink capability.

The transmitter [I] of this thesis achieves a similar link strength as the BLE transmitter in [25] in the high-power mode but consuming 55% less energy per bit. On the other hand, the transmitter [I] achieves a higher link strength than
the BFSK, BPSK and BLE transmitters in [11], [12] and [48] while consuming similar energy per bit. For these reasons, the FOM of [I] is better compared to these works. The FOM of transmitter [I] is in the same range with that of the BPSK transmitter in [45] because the power efficiencies of these transmitters are somewhat equal and because both use energy-efficient modulation schemes that require somewhat similar SNR per bit. 64-DPPM requires a slightly lower SNR per bit than BPSK (see Fig. 3.11) which enables a better uplink performance with respect to the transmitted energy per bit. For this, the FOM of [I] is better than that of [45].

Comparison of transmitters using the new FOM and plotting the link strengths versus the EPBs provides a new perspective on the energy efficiencies of published ULP narrowband transmitters. Comparison of individual metrics such as power consumption, output power, power efficiency or EPB does not provide a very comprehensive and meaningful view of the energy efficiency. The new FOM accounts for multiple transmitter properties that impact the uplink performance, and considers how high the link strength is relative to the EPB. In this way, the FOM reflects the energy-related challenge in radio transmitters: it is difficult to increase the link strength without increasing the EPB. Increasing the link strength requires a greater SNR or a lower SNR requirement. For a greater SNR, a higher output power can be used which increases the EPB through a higher PA power consumption. Alternatively, a lower noise bandwidth can be used which often requires a lower $f_{BB}$, i.e. a greater $T_{BB}$, which results in an EPB increment through increased active time per bit. Optionally, instead of increasing the SNR, an energy-efficient modulation scheme can be used that requires a low SNR per bit for a given error probability. This way, a better error performance can be achieved relative to the amount of transmitted energy per bit which also implies greater uplink range – the output signal may attenuate by a greater factor before reducing to the SNR level required for the targeted error probability. Note that, between different modulation schemes, there can be differences greater than 9 dB in the SNR required per bit (see Fig. 3.11). The most energy-efficient modulation schemes require even 90% less signal energy per bit than the least energy-efficient ones. The choice of modulation scheme can, therefore, have such a profound impact on the overall energy efficiency that it cannot be ignored in the comparison of transmitters.

5.4 DPPM modulator for a UWB transmitter

5.4.1 Overview

This Section 5.4 briefly discusses the EPB reduction achieved in [IV] by using DPPM with the UWB transmitter instead of OOK. Some additional details and discussion are provided in [IV]. Fig. 5.21(a) shows the UWB transmitter...
architecture which is similar to the narrowband transmitters of Sections 5.1 and 5.2. The greatest difference is that the transmitter front-end is a UWB pulse generator (PG). There are only minor differences in the DPPM modulator compared to the modulator discussed in Section 5.1.2. No guard slots are utilized here and therefore the adder is not included in the DPPM modulator on the SYM line between the memory register and the equality comparator. Guard slots would not bring much benefit in this case – the UWB pulse duration is roughly 4 ns while the length of a slot is approximately $T_{BB} = 100$ ns, as determined by the 10-MHz RO, for which the pulses are separated by a gap even without guard slots. The RO design is equal to that of Fig. 5.14 but a frequency of 10 MHz is used, enabled by the use of a higher bias current.

The UWB PG has not been designed by the author of this thesis and is not discussed in this work in detail. For more information about the PG design, see [2]. Fig. 5.21(b) shows an example of a generated UWB pulse. In the figure, the signal contains undesirable oscillations after $t = 6$ ns. The low-frequency oscillations are caused by IC package parasitics and bonding wire cross talk, and the high-frequency oscillations are reflections of the original UWB pulse between the PG and the oscilloscope. This pulse was measured with a packaged chip. With a wirebonded chip and an improved measurement setup, these undesirable oscillations are significantly reduced as shown in [2, Fig. 5].
The UWB transmitter was implemented in a 180 nm CMOS process. Fig. 5.22 shows a micrograph of the implemented blocks. The following measurement results have been presented in [IV]. The supply voltage of all the blocks is 1.2 V. The pulse generator consumes 39 pJ per transmitted pulse and its standby power consumption is 174 nW. The frequency of the RO, i.e. $f_{BB}$, was tuned to 10.04 MHz and it consumes 1.40 $\mu$W of power.

Table 5.4 shows the measured transmitter performance with continuous-mode data transmission using OOK, 16-DPPM, 32-DPPM and 64-DPPM. The OOK-mode power consumption includes the power consumed by the pulse generator and the ring oscillator. In the case of OOK, the power consumption of the OOK modulator is omitted because non-optimal digital logic was used to convert a bit stream (provided by the memory register) to an edge-triggering signal for the pulse generator. The power consumption in DPPM mode includes the power consumed by the pulse generator, the RO and the DPPM modulator. The average data rates with DPPM without guard slots can be calculated with (3.9). The EPB with OOK and the lowest EPB with DPPM are 19.7 and 8.58 pJ/bit, respectively. DPPM thus achieves up to 56% lower EPB than OOK.
5.5 Summary

This chapter presented two narrowband transmitters and a DPPM modulator for a UWB transmitter. The narrowband transmitters operate in the 433-MHz band and can transmit OOK and DPPM modulated data. With data manipulation, also PPM transmission is possible [I]. Both narrowband transmitters achieve a lower EPB using DPPM instead of OOK. One key point was that, despite the lower EPB, the error performance is expected to be better with DPPM because it requires lower $\gamma$ than OOK for the targeted PER of $4.8 \cdot 10^{-4}$.

The first narrowband transmitter is based on the PO-based topology. It consumes only 67 nW of power when transmitting 64-DPPM modulated data in packet-mode at a data rate of 4.8 kbps. In this mode, the EPB is 14 pJ/bit. Despite the ultra-low power consumption and EPB, the data was successfully received at a distance of 30 meters in the measurements without exploiting directional antennas. One highlight in the uplink measurement was that no carrier nor symbol synchronization was required between the DPPM transmitter and the receiver. The second narrowband transmitter is based on the direct-modulation topology. It was designed for a higher output power for a higher uplink range. This TX consumes 1.56 $\mu$W of power when transmitting data at a rate of 2.27 kbps. Thus, the energy consumption per bit is 0.69 nJ/bit. With the FOM of $-189.7$ dBJ/bit, the second transmitter is state of the art in terms of energy efficiency as shown by Table 5.3 and Fig. 5.20. The $-175.6$-dBJ/bit FOM achieved by the PO-based transmitter is not as good because the radiated power is limited by, for instance, the electrically small antenna used in the PO. In addition to the narrowband transmitters, a DPPM modulator was designed for a UWB transmitter. Using 64-DPPM instead of OOK, the EPB of the UWB transmitter is reduced by 56% from 19.7 to 8.58 pJ/bit.
6. Ultra-low power gesture sensor interface

This chapter discusses the 3.2-µW capacitive gesture sensor interface presented in [I]. It was designed for the energy-autonomous wireless sensor SoC which was discussed in Section 1.2. The gesture sensor can be used to detect push and sweep gestures and it enables human interaction with the sensor node. The sensor interface has been fabricated on the same integrated circuit with the narrowband transmitter presented in [I] and Section 5.2 of this thesis. The transmitter could be used for transmitting, for instance, the raw gesture sensor data or other data when triggered by a gesture from a user. The gesture sensor design in [I] is an improved version of the sensor interface presented in [V]. The interface in [V] consumes less power, 462 nW, but its gesture detection range is lower. The discussion in this chapter focuses on the design presented in [I] due to its improved detection range. Only a short sensor design description was provided in [I]. This chapter supplements the description with additional design details and linearity measurement results which were not included in [I].

6.1 Overview

An interest has grown in capacitive proximity and gesture sensors and their applications [V], [53]–[60]. These types of sensors could be used in human interface devices, for instance, in ambient intelligence, wearables and vehicular applications [56], [58], [59]. Some other applications include robot collision avoidance and seat occupancy detection [53], [60]. Capacitive proximity sensing enables touchless gesture control with low cost, light weight and ultra-low power consumption without wearable accessories. Moreover, a capacitive proximity sensor can detect objects over a wide angle. The major drawbacks, however, are limited detection distance and the fact that capacitive proximity sensors are sensitive to dynamic electric fields in the environment of the sensor [56].

Fig. 6.1 shows the basic building blocks of a capacitive proximity sensor. A user’s hand affects the capacitance of the sensing element that can be, for example, a copper plate. The sensor readout circuit converts the capacitance, or the change of the capacitance, to an electrical quantity such as voltage or...
frequency. Furthermore, this quantity is typically converted to digital data for digital signal processing such as proximity or gesture detection. A change of capacitance may be converted to voltage using, for instance, a charge-sensitive amplifier [61]. For capacitance-to-frequency conversion, a capacitively controlled oscillator [53], [62] can be used. Furthermore, voltage-to-digital and frequency-to-digital conversion can be performed using a $\Delta\Sigma$ ADC [63] and a reset counter [62], [64], respectively.

When a single proximity sensor is used, generally only one-dimensional information is obtained about the location of the hand. This limits the versatility of the detectable gestures but, for example, the approach of a hand could be detected in addition to a tap or touch of the sensor plate [3]. By combining two parallel proximity sensors, two-dimensional information is obtained. This enables detection of a wider range of motions such as sweeps from left to right or vice versa in addition to the approach, tap and touch.

In this thesis work, capacitive 2-axis gesture sensor interfaces were implemented. Gesture sensing can also be performed using, for instance, a camera, data glove, accelerometer, thermal imager or ultrasonic tracking. Additionally, radar can be used. Generally, the methods differ in system complexity, accuracy, operational range, power consumption, calibration requirements and the type and rate of data output. More information about these methods can be found, for instance, in references [59], [65] and [66].

6.2  ULP 2-axis gesture sensor interface in 180 nm CMOS

6.2.1  Sensor system

Fig. 6.2 shows the block diagram of the 2-axis gesture sensor presented in [I]. It contains two proximity sensor (PS) interfaces. In each PS interface, a capacitively controlled ring oscillator (CCRO) produces a frequency that is affected by the capacitance $C_S$ of a sensor element. The sensor element, depicted in the bottom left, contains three copper plates in coplanar arrangement [54], [55]. The capacitance $C_S$ is nominally determined by the capacitance between
two copper plates of which other one is grounded. When brought near the plates, a user’s hand increases $C_S$. This decreases the oscillation frequency of the CCRO. With two parallel proximity sensors, information is obtained about both the horizontal and vertical position of the hand. The grounded middle plate is shared by the two PS interfaces.

Fig. 6.3 shows the schematic of the CCRO. It is a current-starved three-stage ring oscillator followed by two buffering inverters. With current-starving, the power consumption of the CCRO can be limited which, however, also limits the oscillation frequency. The bias current $I_B$ is provided by an on-chip bias current generator. The sensor element capacitance $C_S$ is connected as a load for the inverter INV$_1$. The two other inverters, INV$_2$ and INV$_3$, are loaded by the metal-insulator-metal capacitors $C_2$ and $C_3$, 400 fF each, to decrease the phase noise. The output of the CCRO, denoted with $f_{ccro}$, is fed to a modified cascaded integrator-comb (CIC) filter [V]. This filter converts the time-domain input signal to 24-bit digital data, clocked at a lower frequency. The filter operation is discussed in more detail in Section 6.2.2. As shown in Fig. 6.2, the CIC filter is followed by a digital adder/subtractor and a multiplier. With programmable offset and gain values, the offsets of the two PS interfaces can be corrected and gains can be adjusted. The digital 24-bit outputs of the left and right PS interface ($D_L$ and $D_R$ in Fig. 6.2) are provided to the memory register.
From the memory register, they can be accessed through the SPI by an external controller such as a computer, a microcontroller or an FPGA.

The frequency divider in Fig. 6.2 divides the frequency of the reference RO output, CLK\textsubscript{ref}, to produce the clock signals CLK\textsubscript{D1} and CLK\textsubscript{D2} for the modified CIC filters. The reference RO is a current-starved three-stage ring oscillator similar to the CCRO of Fig. 6.3. As opposed to the CCRO, all inverters INV\textsubscript{1}–INV\textsubscript{3} of the reference RO are each loaded by a 31-fF MIM capacitor and INV\textsubscript{1} is not loaded by any sensing capacitance. The reference RO was initially\textsuperscript{1} designed to produce an oscillation frequency in the range of 50 kHz while consuming a couple of tens of nanowatts of power and supplied with 0.8 V. The frequency of CLK\textsubscript{ref} is divided by a series of 30 divide-by-two frequency dividers. The clocks CLK\textsubscript{D1} and CLK\textsubscript{D2} can be multiplexed from the outputs of the divide-by-two circuits. The sample rate in the CIC filter output is \( f_{s,CIC} = \frac{f_{ref}}{R_1 \cdot R_2} \), where \( R_1 \) and \( R_2 \) are programmable frequency division ratios.

### 6.2.2 Modified decimating CIC filter

Fig. 6.4 shows the block diagram of the modified decimating CIC filter. The filter has been discussed briefly in [V] and [I]. The following depiction of the operation contains additional details related to the used synchronization scheme.

A conventional CIC filter [67], [68] is fed with 1-bit or multibit data clocked with a constant frequency. In the modified filter of Fig. 6.4, the data lies in the frequency of the input signal and the input signal is fed to the clock input of the first integrator instead of the data input. Synchronizers have been added to reduce the probability of synchronization errors that could occur as data is

\textsuperscript{1}The reference RO was originally designed for the earlier gesture sensor interface prototype of [V] which uses a 0.8-V supply. However, here we discuss the later gesture sensor prototype [I] where the same reference RO is used with a supply of 0.9 V – the same voltage level that the on-chip direct-modulation transmitter uses.
moved from the domain of clock $f_{ccro}$ to the domain of clock $CLK_{ref}$, i.e. from the first integrator to the second integrator and the downsampler. This is explained in more detail shortly. Nonetheless, the filter low-pass filters the input signal and performs downsampling in a similar fashion to a conventional decimating CIC filter.

The input to the filter is $f_{ccro}$, the output of a CCRO, which is fed to the clock input of the first integrator. The data input is a constant bit 1. The output of the integrator, signal $A_1$, is 24-bit data that is incremented by one at the rising edges of $f_{ccro}$. The second integrator is clocked by $CLK_1$ which is a low-frequency clock, derived by sampling low-frequency clock $CLK_{D1}$ at the falling edges of $f_{ccro}$ in the synchronizer SYNC1. $CLK_{D1}$ is derived from $CLK_{ref}$ by dividing its frequency by the frequency division ratio $R_1$. The output of the second integrator, signal $A_2$, is downsampled by the register at the bottom middle, clocked by $CLK_2$. $CLK_2$ is also derived from $CLK_{ref}$ by frequency division and by using synchronization. The output of the downsampler feeds the first comb whose output furthermore feeds the second comb. The combs are clocked with $CLK_2$ and the output of the second comb, DATA_OUT, is the output of the modified CIC filter. Thus, the filter output sample rate is equal to the frequency of $CLK_2$.

In the integrator section, data moves from the domain of $f_{ccro}$ to the domain of $CLK_{ref}$ from which the clocks $CLK_{D1}$ and $CLK_{D2}$ are derived from. $CLK_{ref}$, $CLK_{D1}$ and $CLK_{D2}$ are asynchronous with respect to $f_{ccro}$ for which errors arise due to timing issues if synchronization is not used. A problematic situation is depicted in Fig. 6.5. When a rising edge of $f_{ccro}$ triggers REG1, the 24 parallel binary signals in the adder outputs $A_1$ and $A_2$ could toggle states for several hundreds\(^2\) of picoseconds. If REG2 was clocked directly with $CLK_{D1}$, the signal $A_2$ would occasionally be triggered to REG2 right when $A_1$ is toggling which also forces $A_2$ to toggle. In Fig. 6.5, this occurs at the time instant $t_1$. As a consequence, an erroneous value would be triggered to REG2. Generally, this could occur whenever the signal $f_{ccro}$ rises near the rising edge of $CLK_{D1}$ if $CLK_{D1}$ was directly used to trigger the data in $A_2$ to REG2. This type of error can be avoided by implementing synchronization.

To reduce the probability of timing errors in this work, two-flip-flop synchro-

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\(^2\)This toggling time depends on the propagation delays of on-chip data and clock signals that are affected, for example, by process corner, parasitic capacitances of the signal lines and the timing constraints utilized in the place-and-route tool.
Ultra-low power gesture sensor interface

![Micrograph of the implemented blocks](a)

![Copper sensor plate arrangement](b)

**Figure 6.6.** (a) Micrograph of the implemented blocks, and (b) copper sensor plate arrangement utilized in the measurements. Subfigure (a) adapted from [I], used under CC BY 4.0 / TX-related notation changed to grey. Subfigure (b) from [V], © 2017 IEEE.

nizers [69] SYNC1 and SYNC2 have been added. SYNC1 samples CLK$_{D1}$ at the falling edges of $f_{ccro}$ and produces the clock signal CLK$_1$. As SYNC1 consists of two flip-flops in series, CLK$_1$ toggles at the second falling $f_{ccro}$ edge following the transition of CLK$_{D1}$, at $t_2$ in Fig. 6.5. REG2 is clocked with CLK$_1$ and, thus, the signal $A_2$ is triggered to REG2 closer to a falling edge of $f_{ccro}$ instead of its rising edge. In SYNC1, the first flip-flop could go metastable if the flip-flop data input (CLK$_{D1}$) toggles near the time instant when the flip-flop receives a triggering clock edge (inverted $f_{ccro}$ rises) [69]. The output of the first flip-flop is sampled to the second flip-flop one $f_{ccro}$ cycle later for which the probability of CLK$_1$ going metastable is significantly reduced [69]. SYNC2 operates with the same principles as SYNC1 and synchronizes CLK$_{D2}$ to the falling edges of $f_{ccro}$.

### 6.2.3 Measurement results

The gesture sensor interface of publication [I] was implemented in a 180 nm CMOS process. It is included on the same chip with the direct-modulation narrowband DPPM transmitter of Section 5.2, also published in [I]. The gesture-sensor-related blocks are denoted in the micrograph of Fig. 6.6(a) on the left. Fig. 6.6(b) shows the sensor element PCB utilized in the measurements. The size of each of the three copper plates is 4cm-by-7cm and they are separated by a gap of 1.5 cm. The backside of the sensor element PCB is devoid of copper. The
plates were connected to the measurement PCB\(^3\) with jump wires. The middle plate was connected to the measurement PCB ground.

The main measurement results have been presented in [I]. As explained in [I], a supply voltage of 0.9 V was used with the reference RO, CCROs and DSP block. The frequencies of the left and right side CCROs were set to the maximum values of 269 and 219 kHz, respectively, for a maximum SNR. The reference RO was set to a frequency of 48.4 kHz. The clock division ratios \(R_1\) and \(R_2\) were set for a total division ratio of 1024. This results in a sample rate of 47.2 SPS at the CIC filter output which is fast enough for detecting hand sweep and push gestures. In this configuration, the CCROs consume 1.43 and 1.40 µW of power. The reference RO and the DSP block consume 13 nW and 0.40 µW, respectively. Thus, the 2-axis gesture sensor interface consumes a total of 3.24 µW. The BCGs for the reference RO and CCROs were not optimized in terms of power consumption and their power was not included in the results in [I].

The linearity of the sensor output was not presented nor discussed in [I] but was measured as follows. The static outputs of the two PS interfaces were characterized in the style of comparable PS works of [53] and [54]. A grounded copper plate was placed on top of the sensing copper plate, i.e. one of the outmost plates in Fig. 6.6(b), that was connected to the PS interface under study. The size of the grounded plate was 3cm-by-4cm. The grounded plate was placed at different distances \(d\) from the sensing plate and the CCRO frequency and digital PS output were observed and recorded. The measurement was performed individually for the left and right PS interface. Fig. 6.7(a) shows the frequency change of the left and right CCROs in percents versus distance \(d\). The frequency change here refers to the change compared to the nominal CCRO frequency without the grounded test plate near the sensor element. Fig. 6.7(b) shows the change of the digital output. The frequency change is relatively small, less than 1%, if the distance is greater than 5 cm. The digital output is practically a linear

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\(^3\)The measurement PCB here refers to the PCB containing the sensor interface microchip and other components related to the measurements.
function of the frequency and, hence, its magnitude of change is comparable to that of the frequency.

The static output of the left side interface was also measured using the measurer’s palm as the test object. To maintain a roughly fixed distance between the sensor plate and the palm, two plastic rods were placed on top of the sensor element PCB at distance $d$. During the measurement, the palm was held in place, resting on the rods. The measured change of the digital output versus $d$ is shown in Fig. 6.7(b). The presence of the palm produces a slightly greater amplitude change compared to the grounded 3cm-by-4cm copper plate. A palm at a distance of 12 cm from the sensor plate changes the digital output by roughly $-0.28\%$.

To test the dynamic performance of the sensor interface, hand sweep and push gestures were performed and the digital outputs of the left and right PS interfaces (i.e. $D_R$ and $D_L$ in Fig. 6.2) were recorded. Also in this measurement, the plastic rods were used to limit the distance between the palm and the sensor element PCB. During sweeps, the palm was swept along the surface of the rods. During push gestures, the palm approached the sensor plate and was stopped by the rods. Fig. 6.8, adapted from [I], shows the differential output, $D_R - D_L$, as the gestures were performed with the distance limited to 7 cm and 12 cm. The sequence of gestures was performed in this order: two sweeps from left to right, two sweeps from right to left, two pushes over the left plate and two pushes over the right plate. With the palm over the left plate, the differential output increases. With the palm over the right plate, it decreases. With $d \approx 12$ cm, the peaks produced in the differential output vary roughly from 80 to 175 LSBs. With a shorter distance, the peaks are higher. The measured root mean square (RMS) noise in the differential output is approximately 15.9 LSBs.

It can be seen in Fig. 6.7 that the PS outputs are highly nonlinear. This may make it challenging to calculate the absolute location of the user’s hand. However, for detecting basic push and sweep gestures, the nonlinearity is not...
expected to be a major deficiency. With the sweep and push gestures clearly recognizable in Fig. 6.8, it is fair to assume that the gestures could be detected with some accuracy using some detection algorithm. However, it is difficult to predict the detection accuracy without further tests. Such tests are out of the scope of this thesis as the main focus is on energy-efficient ULP transmitters.

This gesture sensor is on the same IC with the transmitter which was presented in Section 5.2 and [1]. The sensor consumes 3.24 µW of power when producing two 24-bit data streams at the sample rate of approx. 47.2 SPS. The transmitter, on the other hand, consumes 1.56 µW of power when it is configured to transmit 48-bit data packets at the sensor output sample rate, 47.2 packets/second (see Section 5.2.4). Thus, if the transmitter was used to transmit the raw sensor data, together they would consume 4.8 µW. As a photovoltaic array with an area of 6.5 cm² can produce 25 µW under office illumination [2], the sensor and transmitter could be powered using a reasonably sized energy harvester.

6.3 Summary

Capacitive gesture sensing offers a way to implement human interface devices for ULP applications. It enables sensor power consumptions in the range of microwatts. A 2-axis gesture sensor interface was implemented that is based on capacitively controlled ring oscillators. It consumes 3.24 µW of power and the sample rate of the digital outputs is 47 SPS. Hand sweeps and push gestures performed at a distance of 12 cm from the sensor plates are visually recognizable in the output data. Additional tests using a gesture detection algorithm would be required to assess the level of gesture detection accuracy enabled by the presented sensor interface.
7. Summary and conclusion

This doctoral thesis focused mainly on ULP narrowband radio transmitters. Prior ULP narrowband transmitters that have been capable of sub-mW power consumption have mostly used OOK, BFSK and BPSK. These binary modulation schemes enable low power consumption as they can be implemented without an excessive amount of high-power circuit blocks. Moreover, they require a relatively low SNR per bit to achieve a given error probability which enables reduced output power and, thus, reduced PA power consumption.

In contrast to the majority of the earlier works, this thesis investigated two optional modulation schemes that could be used in ULP narrowband transmitters instead of OOK, BFSK and BPSK: PPM and DPPM. Particularly, the focus has been on their M-ary implementations. With M-ary encoding, multiple bits are encoded per transmitted RF waveform and the carrier is transmitted for a reduced time per bit. The consequent reduced active time per bit may translate to a reduced energy consumption per bit. However, modulation schemes require different amounts of SNR per bit to achieve a given error probability. Thus, when considering different modulation options, one must also consider how the choice of modulation scheme affects the required output power. The reduced active time per bit does not necessarily bring benefits in terms of energy efficiency if the modulation scheme also requires significantly greater output power for the desirable error performance.

M-ary PPM and DPPM both require a low SNR per bit. With PPM, this is enabled by signal orthogonality. DPPM is not an orthogonal modulation but enables the use of demodulation concepts similar to those used with PPM. In Section 3.2.4 and [II], a packet-level soft-decision decoding method was described for receiving a DPPM signal that requires a low SNR per bit to achieve a decent packet error ratio. Due to the low required SNR per bit and the low active ratio, PPM and DPPM are considerable options for wireless devices for which energy-efficient transmission is favorable.

In Chapter 4 and [I], PPM and DPPM were compared with OOK, BPSK and BFSK in a new way. The new comparison method considered how the choice of modulation scheme affects the combined energy consumed per bit by the power amplifier and carrier synthesizer when, with each modulation scheme,
the output power is scaled for equal error performance. The results suggested that the 16-ary and 64-ary variants of PPM and DPPM are more energy-efficient than OOK, BPSK and BFSK. The disadvantages with the M-ary PPM and DPPM schemes are, however, the lower data rate and longer transmit time with a given baseband clock frequency compared to the binary modulation schemes. The longer transmit time can cause a higher EPB at a receiver. For this, PPM and DPPM may be more suitable for systems where ultra-low power devices communicate with a device that has an energy source with a greater capacity. Moreover, a PL-SDD DPPM receiver is more complex than a basic OOK, BPSK or BFSK receiver as it requires memory and more complex DSP.

As a part of this thesis work, two ULP narrowband DPPM transmitters were implemented for the 434-MHz band. The first one, originally presented in [II], is a PO-based transmitter and its peak output power is –25 dBm. This transmitter consumes energy down to 11.6 pJ/bit. When configured to transmit data in 48-bit packets at the rate of 100 packets per second (i.e. 4.8 kbps), the transmitter consumes only 67 nW of power. Despite the ultra-low EPB and power consumption, the data transmitted in this mode was received at a distance of 30 meters from the transmitter using a custom-made PL-SDD DPPM receiver and without exploiting highly directional antennas. The second transmitter, originally presented in [I], is based on the direct-modulation architecture and its peak output power is –2.1 dBm. When configured to transmit 64-DPPM data in 48-bit packets and 47.2 packets per second (i.e. 2.27 kbps) with the timing reference symbol, this TX consumes 1.56 µW and 0.69 nJ/bit. It was estimated that the line-of-sight uplink range is up to one kilometer. The range is notable considering the ultra-low power consumption and EPB. In continuous-mode 64-DPPM transmission, the measured average power consumptions of the PO-based and direct-modulation transmitter are 10.4 and 41.1 µW, respectively. In OOK mode, the presented transmitters can transmit data at a higher data rate but also consume more power and more energy per bit.

The numerous published ULP and other transmitters use a variety of different modulation schemes and also differ significantly in terms of the power consumption, output power, data rate and EPB. Due to all the differences, comparing these transmitters and their energy efficiencies is not a straightforward task. The transmitter energy efficiency FOMs that have been used previously have generally neglected the noise bandwidth and the SNR requirement of the modulation scheme. However, it was brought up in [I] that these parameters have a major impact on the uplink performance and should be accounted for in the FOM. The noise bandwidth affects the maximum SNR achievable with the transmitted signal. Furthermore, this maximum achievable SNR and the SNR required by the modulation scheme together determine how much the output signal may attenuate before falling to the level required for the desirable error probability. To achieve a certain error probability such as BER = 10^{-5}, the most energy-efficient modulation schemes require less than 10% of the SNR that the least energy-efficient schemes require per bit. Due to this, the SNR
requirement of the utilized modulation scheme has a profound effect on the uplink capability and the overall energy efficiency of a transmitter. Therefore, the SNR requirement should not be disregarded when the energy efficiencies of transmitters are compared.

In Section 2.4 and [I], a new FOM was derived for transmitter energy efficiency that accounts also for the noise bandwidth and the SNR required by the modulation scheme. This new FOM provides a rational metric for the energy efficiency and a new perspective on radio transmitters. By referencing the link strength to the EPB, the FOM reflects the fundamental energy-related challenge in transmitters – it is difficult to achieve a high link strength consuming low energy per bit. It was concluded in Section 5.3 and [I] that the keys to good energy efficiency are the use of power-efficient circuitry and the use of a modulation scheme that requires a low SNR per bit. The FOM was calculated for the narrowband transmitters of this work and a multitude of state-of-the-art ULP transmitters. The results in Table 5.3 and Fig. 5.20 show that the DPPM transmitter presented in [I] is indeed state of the art in terms of energy efficiency. The FOM calculation is based on parameters that are generally reported in transmitter publications. Therefore, it could be used for comparing a wide range of published transmitters.

In addition to the radio transmitter discussion and designs, this thesis discussed the capacitive 2-axis gesture sensor interfaces of [I] and [V]. The interface in [I] was designed for the energy-autonomous SoC which was discussed in Section 1.2 and the sensor was included on the same integrated circuit with the transmitter of Section 5.2, also published in [I]. The gesture sensor interface of [I] provides two digital outputs that enable detection of hand sweeps from left to right and vice versa in addition to push gestures. The power consumption is 3.24 μW. The measured results show that, with the distance between the palm and the sensor plates limited to 12 cm, the performed gestures are visually distinguishable in the sensor output data. For a more comprehensive analysis, a gesture detection algorithm should be used to check how accurately the gestures can be recognized with this kind of sensor circuit performance. The 3.24-μW gesture sensor interface is based on the earlier interface presented in [V]. The interface in [V] consumes only 462 nW of power but also the detection range is significantly reduced. For that, the sensor discussion focused on the work presented in [I].

The gesture sensor part of the thesis can be considered as an experimental study of what kind of performance can be achieved with μW-range power consumption using capacitively controlled ring oscillators. The results are provided as they are and, due to the extensive work required by the transmitter design and the related energy efficiency considerations, no attempt was made to provide an in-depth analysis of the sensor systems.
A. Appendix: Derivation of PER of DPPM with packet-level soft-decision decoding (PL-SDD) scheme

In [II], a short version was shown of the derivation of the PER of DPPM with the PL-SDD scheme in the AWGN channel. The following derivation of this PER contains additional steps and details compared to the derivation in [II].

In this PER analysis, let us assume that the PL-SDD DPPM receiver stores \( L_{p,max} = 1 + N_s \cdot (2^B + 1) \) samples of the envelope detector (ED) output\(^1\). \( N_s \) is the number of symbols included in a packet. The distribution of the ED output in the “on” and “off” slots is given by (3.21) and (3.22), respectively. With \( N_s \) symbols per packet, the DPPM receiver knows that, with the start pulse, \( N_O = N_s + 1 \) “on” slots are contained in \( L_{p,max} \) slots. The receiver interprets the \( N_O \) slots with the highest amplitudes as the “on” slots. The rest are interpreted as “off” slots. This operating principle was depicted in Fig. 3.9 with \( N_s = 2 \) and \( B = 3 \) in which case the receiver analyzes \( L_{p,max} = 19 \) samples of the ED output.

The PER can be derived with probability calculations. Let \( O \) and \( Z \) be arrays containing the ED output amplitudes corresponding to the “on” and “off” slots, respectively. There are \( N_O \) “on” slots in a packet, i.e. within \( L_{p,max} \) slots, and hence the number of elements in \( O \) is \( N_O \). The rest of the slots are “off” slots and the number of elements in \( Z \) is \( N_Z = L_{p,max} - N_O \). A packet error occurs if at least one element in \( O \) is lower than the highest element in \( Z \) in which case at least one “off” slot is interpreted as an “on” slot. The PER is the probability of this error and, as given in [II], it can be written as

\[
PER_{dppm} = \int_0^{\infty} P(\text{max } Z = r) \cdot P(\text{min } O \leq r) \, dr, \tag{A.1}
\]

where \( P(\text{max } Z = r) \) is the probability density function (PDF) of the maximum value in \( Z \) and \( P(\text{min } O \leq r) \) is the probability that the minimum value in \( O \) is

\(^1\)If \( N_s, B \) or both are high, \( L_{p,max} \) is high and plenty of memory is required for storing all \( L_{p,max} \) ED output amplitudes. However, only the timing of the \( N_s+1 \) highest peaks is required for successful decoding. Thus, for lower memory usage, such an algorithm could possibly be implemented that stores less than \( L_{p,max} \) ED output samples with timing information.
less than or equal to \( r \). \( P(\max Z = r) \) is given by [II]

\[
P(\max Z = r) = \frac{d}{dr} \left[ \int_0^r \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \, dr \right]^{N_Z}.
\]  \tag{A.2}

The meaning of (A.2) is explained as follows. The integral in the square brackets in (A.2) is the probability that the ED output amplitude corresponding to a single “off” slot is less than or equal to \( r \). This integral is a cumulative distribution function (CDF). Raised to the power of \( N_Z \), this CDF gives the probability that all the \( N_Z \) elements in \( Z \) are less than or equal to \( r \). The PDF of a continuous random variable is obtained by differentiating its CDF. Thus, the derivative of the CDF raised to the power of \( N_Z \), i.e. (A.2) as a whole, gives the PDF of the maximum value in \( Z \). Substituting (3.22) to (A.2) yields

\[
P(\max Z = r) = \frac{N_Z r}{\sigma^2} e^{-r^2/2\sigma^2} \left[ \int_0^r \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \, dr \right]^{N_Z-1}.
\]  \tag{A.3}

The derivative in (A.3) can be solved using the chain rule which states that the derivative of \( f(g(x)) \) is \( f'(g(x)) \cdot g'(x) \). This yields

\[
P(\max Z = r) = \frac{N_Z r}{\sigma^2} e^{-r^2/2\sigma^2} \left[ \int_0^r \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \, dr \right]^{N_Z-1}.
\]  \tag{A.4}

In (A.1), \( P(\min O \leq r) \) is given by [II]

\[
P(\min O \leq r) = 1 - \left[ 1 - \int_0^r p_1(r) \, dr \right]^{N_O}.
\]  \tag{A.5}

The content of the square brackets in (A.5) is the probability that the ED output in a single “on” slot is greater than \( r \). Raised to the power of \( N_O \), this gives the probability that all the \( N_O \) elements in \( O \) are greater than \( r \). Subtracting this from 1, i.e. (A.5) as a whole, then gives the probability that at least one element in \( O \) is less than or equal to \( r \). Substituting (3.21) to (A.5) yields

\[
P(\min O \leq r) = 1 - \left[ 1 - \int_0^r \frac{r}{\sigma^2} e^{-(r^2+A^2)/(2\sigma^2)} I_0 \left( \frac{rA}{\sigma^2} \right) \, dr \right]^{N_O}.
\]  \tag{A.6}

Finally, substituting (A.4) and (A.6) to (A.1) yields

\[
\text{PER}_{dppm} = \int_0^\infty \frac{N_Z r}{\sigma^2} e^{-r^2/2\sigma^2} \left[ \int_0^r \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \, dr \right]^{N_Z-1} \cdot \left\{ 1 - \left[ 1 - \int_0^r \frac{r}{\sigma^2} e^{-(r^2+A^2)/(2\sigma^2)} I_0 \left( \frac{rA}{\sigma^2} \right) \, dr \right]^{N_O} \right\} \, dr.
\]  \tag{A.7}

This is the packet error ratio with DPPM with the packet-level soft-decision decoding, given in (3.23) and [II]. The function can be evaluated, for instance, using numerical integration in mathematical software. The waveform-level PER simulation results in Section 3.3 with 16-DPPM and 64-DPPM using 48-bit packets suggest that the equation is correct (see Fig. 3.13).
References


[43] “Short Range Devices (SRD) operating in the frequency range 25 MHz to 1 000 MHz; Part 2: Harmonised Standard for access to radio spectrum for non specific radio equipment,” ETSI EN 300 220-2, Version 3.2.1, European Telecommunications Standards Institute, Jun. 2018.


Errata

Publication I

Correction 1: On page 1773, the journal article says: “E, E_b and N_0 are the energy of the signal per symbol, the energy of the signal per bit and noise power spectral density, respectively.” The quoted sentence is correct regarding E_b and N_0 but not regarding the energy E. The sentence should be replaced with: “E is the energy conveyed by the unmodulated carrier wave over a time interval whose length is T_{BB}. E_b and N_0 are the energy of the signal per bit and noise power spectral density, respectively.” The discussed energy E is not always equal to symbol energy.

Correction 2: On page 1776, the article says: “...where E_{sig} and E_n are the RF carrier energy per symbol and the noise energy in the noise BW per time interval T_{BB}, respectively.” The sentence is correct regarding E_n but not regarding E_{sig}. This sentence should be replaced with: “...where E_{sig} is the energy conveyed by the unmodulated carrier wave over a time slot whose length is T_{BB}, and E_n is the energy of the noise within the noise bandwidth, also in a time slot the length of T_{BB}.”

Correction 3: On page 1779 of the article, it is said that: “If higher data rate is desirable, the use of quadrature or 8-ary PPM or DPPM could be considered.” The word “quadrature” in the sentence should be “quaternary”.

Publication IV

The conference paper discusses EPB as if it was synonymous with energy efficiency. However, it is concluded in Sections 2.3 and 2.4 of this thesis that the EPB is not a proper measure for the energy efficiency. For that, the paper rather discusses EPB reduction, not energy efficiency improvement. EPB reduction can be beneficial but does not directly mean that energy efficiency is improved.