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Title: Discussion on design of WUS with OFDM based sequence
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1 Introduction

In this contribution, we discuss sequence-based WUS design. We first give a brief summary of the sequence-based WUS in Rel-19 before proposing potential avenues for a 6G WUS.

In RAN1#122-bis, the following agreement was reached

> Agreement: **RAN1#122-bis**
>
> Study and evaluate DL WUS of OFDM based sequence and corresponding mechanisms for 6GR EE improvement, regarding at least the following aspects:
>
> - Coverage target for DL WUS (e.g., same as PDCCH, common sync signal, or other)
> - Measurements and/or synchronization.
> - System overhead and network energy consumption/UE energy saving for UE operation with the DL WUS.
> - RRC states
> - Other functionalities

2 Overview of Rel-19 Sequence-based LP-WUS Design

In this section, we briefly review the sequence-based LP-WUS design in Rel-19 which will serve as a baseline scheme in our evaluations.

2.1 OOK Encoding

Consider B information bits $\boldsymbol{b} = [b_0, b_1, \dots, b_{B-1}]$ transmitted in L OFDM symbols over N_{SC}^{WUS} sub-carriers. There are $M \in \{1,2,4\}$ OOK time-domain symbols per OFDM symbol. The encoding procedure consists of three steps:

1. Channel Coding
2. Rate-Matching (RM)
3. Line Coding (Manchester Coding)

Channel coding uses the NR small-block length coding schemes and RM will truncate the encoder output to $E = G/2$ bits with $G = LM$ denoting the total number of OOK symbols of the WUS. Line coding will map one input bit to two output bits by applying Manchester coder resulting in G encoded bits. This part is referred to as OOK encoding in the example is given in Figure 1.

A pure sequence-based WUS design will not use this OOK encoding scheme. The channel coding is used in the OOK encoding to meet the requirements on False-Alarm Rate (FAR) and Missed-Detection Rate (MDR) and is not used in the sequence encoding as subsequently described.

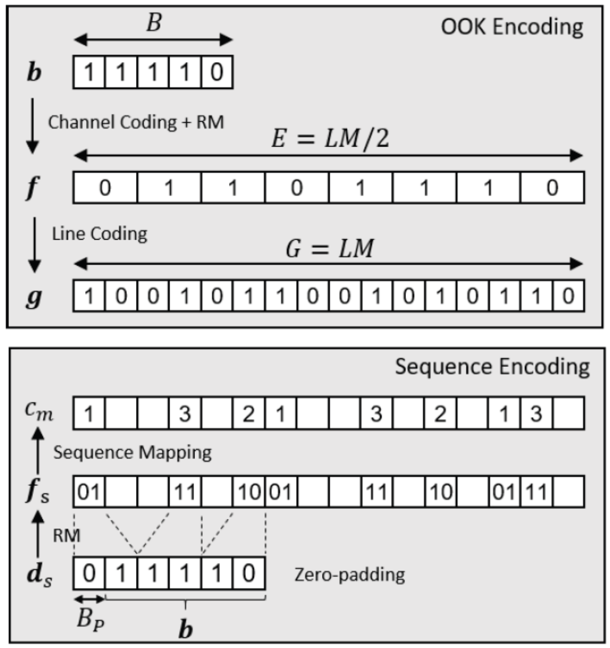

Figure 1: Example of WUS encoding with $\mathrm{B} = 5, L = 4$*,* $M = 4$ *and* $N_{seq} = 4$*.*

2.2 Sequence Encoding

The NR sequences in TS 38.211 are reused (known as cyclically extended Zadoff-Chu (ZC) sequences). The c^{th} ON-sequence $\boldsymbol{r}_c$ of length M_{ZC} is given by

$$\boldsymbol{r}_c(n) = \boldsymbol{x}_q([n + n_{cs}]\ mod\ N_{ZC})$$
$$\boldsymbol{x}_q(i) = e^{-j\frac{\pi q i(i+1)}{N_{ZC}}}$$

with N_{ZC} denoting the largest prime s.t. $N_{ZC} < M_{ZC}$, $q \in \{1,2,\dots,N_{ZC}-1\}$ is the root and n_{cs} is the cyclic shift of the ZC sequence, $n = 0,1,\dots,M_{ZC}-1$ and $i = 0,1,\dots,N_{ZC}-1$. Hence sequence r_c is determined by root q and cyclic shift n_{cs}. To achieve the best correlation performance, the sequences are chosen such that the offset between adjacent cyclic shift values is maximized, i.e.

$$n_{cs} = (c\ mod\ P)\left\lfloor\frac{N_{ZC}}{P}\right\rfloor$$

with $c = 0,1,\dots,N_{seq}-1$ and $P = N_{seq}/N_{root}$ where $N_{root} \in \{1,2\}$ is the number of configurable roots.

Different sequences $\boldsymbol{r}_c$ are used to *directly* encode the information bits $\boldsymbol{b}$**.** The maximum number of sequences allowed N_{seq}^{max}depends on M as $N_{seq}^{max} = \{16,8,4\}$ for $M \in \{1,2,4\}$, respectively. The reason is that a longer sequence (smaller values of M) supports more sequences with good correlation properties than a short sequence. The configured number of sequences $N_{seq} \in \{2,4,8,16\} \leq N_{seq}^{max}$ can encode $\delta = \log_2 N_{seq}$ bits per OOK ON-symbol. Therefore, $\lceil B/\delta \rceil$ ON-symbols are required to encode the payload. Hence, the coded bits $\boldsymbol{d}_s = [d_{s,0}, d_{s,1}, \dots, d_{s,N_s-1}]$ of length $N_s = B + B_p$ are obtained by prepending $B_p = (-B \ mod \ \delta)$ zeros, i.e. $\boldsymbol{d}_s = [\boldsymbol{0}, \boldsymbol{b}]$. Note that zeros are added before the MSB so that both $\boldsymbol{d}_s$ and $\boldsymbol{b}$ still encode the same subgroup IDs. Subsequently, rate-matching is repeating $\boldsymbol{d}_s$ to obtain $\boldsymbol{f}_s$ as

$$f_{s,i} = d_{s,i \ mod \ N_s}$$

with $i = 0,1, \dots, E_s - 1$ where $E_s = E\delta$ and $E = LM/2$ is the number of OOK ON symbols available. In the next step, $\boldsymbol{f}_s$ is segmented into blocks of δ bits such that$\boldsymbol{f}_s = [\boldsymbol{f}_{s,0}, \boldsymbol{f}_{s,1}, \dots, \boldsymbol{f}_{s,E-1}]$ where block $m = 0,1, \dots, E - 1$ is given by

$$\boldsymbol{f}_{s,m} = [f_{s,\delta m}, f_{s,\delta m+1}, \cdots, f_{s,\delta(m+1)-1}]$$

Each block $\boldsymbol{f}_{s,m}$ is encoded with sequence index $c_m \in \{0,1, \dots, N_{seq} - 1\}$ as

$$c_m = (\boldsymbol{f}_{s,m})_{(10)}$$

where $(\boldsymbol{b})_{(10)}$ converts the binary sequence $\boldsymbol{b} = [b_0, b_1, \dots]$ into its decimal representation with b_0 as MSB. Therefore, the sequence $\boldsymbol{r}$ of OOK ON-symbol m is given by $\boldsymbol{r}_{c_m}$. An example of the encoding process for payload $\boldsymbol{b} = [11110]$ is shown in Figure 1.

3 Potential WUS Design

We propose to use a product-code, where the B input bits are encoded *independently* in frequency-domain (vertical) and time-domain (horizontal). This vertical and horizontal coding (VHC) strategy allows the two components to be decoded independently which reduces the complexity in the receiver.

More precisely, separate the input bits $B = B_0 + B_1$ into B_0 and B_1 bits associated with the frequency and time dimension, respectively. This is split is *optional* but can provide two levels of error protection (e.g. high-level grouping info on B_0 and refinement on B_1), where by the B_0 bits can be decoded with lower error probability than the B_1 bits.

Without loss of generality, the transmit message $m \in \{0,1, \dots, M - 1\}$, $M = 2^B$, is given by $m = m_0 + m_1 B_0$ with $M_0 = 2^{B_0}$ and $M_1 = 2^{B_1}$. The transmit signal $\boldsymbol{R}_m \in \mathbb{C}^{Q \times L}$ of message m for Q sub-carriers and L OFDM symbols is given by

$$\boldsymbol{R}_m = \boldsymbol{F}_{m_0} \mathrm{diag}(\boldsymbol{w}_{m_1}),$$

where $\boldsymbol{F}_{m_0} \in \mathcal{F}^{Q \times L} = \{\boldsymbol{F}_0, \boldsymbol{F}_1, \dots, \boldsymbol{F}_{M_0-1}\}$ is the code sequence m_0 in frequency domain and $\boldsymbol{w}_{m_1}$is codeword m_1 of time-domain code $\boldsymbol{W} = [\boldsymbol{w}_0, \boldsymbol{w}_1, \dots, \boldsymbol{w}_{M_1-1}] \in \mathbb{C}^{L \times M_1}$.

The design of codes $\mathcal{F}$ and $\boldsymbol{W}$ depend on the transmission requirements and the number of available dimensions. For instance, $\mathcal{F}$ could be designed for good cross-correlation properties. Another criterion is to limit the number of non-zero elements in $\mathcal{F}$ to reduce both decoding complexity and the effects of multi-path propagation.

The time-domain code (or outer code) $\boldsymbol{W}$ can be orthogonal if $QL \geq M$ or non-orthogonal if $QL < M$. In the non-orthogonal case, a non-coherent code can be used and the B_1 bits $\boldsymbol{d} = [d_0, d_1, \ldots, d_{B_1-1}]$ are encoded as

$$\boldsymbol{c} = \boldsymbol{dG}$$

where $\boldsymbol{G}$ is the generator matrix and $\boldsymbol{c}$ are the coded bits. Subsequently, the modulated N-PSK symbol $w_{m_1 l}$ of message m_1 and symbol l is obtained by

$$w_{m_1 l} = e^{i2\pi c_n/N}$$

where c_n is the n^{th} entry of $\boldsymbol{c}$. Finally, the transmitted sequence $\boldsymbol{r}_{ml}$ is given by

$$\boldsymbol{r}_{ml} = \boldsymbol{f}_{m_0 l} \cdot w_{m_1 l}$$

where $\boldsymbol{f}_{m_0 l}$ is the sequence on symbol l corresponding to message m_0. As an example, consider $N = 4$ (e.g. QPSK), $B_1 = 8$ and $L = 14$, a good generator matrix in GF4 is given by

$$\boldsymbol{G} = \begin{bmatrix} 1 & 0 & 0 & 0 & 1 & 3 & 3 & 0 & 1 & 2 & 0 & 0 & 1 & 2 \\ 0 & 1 & 0 & 0 & 1 & 3 & 3 & 0 & 2 & 0 & 1 & 2 & 0 & 1 \\ 0 & 0 & 1 & 0 & 3 & 0 & 2 & 0 & 1 & 1 & 1 & 0 & 2 & 3 \\ 0 & 0 & 0 & 1 & 1 & 0 & 2 & 0 & 1 & 0 & 2 & 2 & 1 & 3 \end{bmatrix}$$

Before evaluating the different transmission schemes, we will discuss receive algorithms and theirs associated complexity.

4 Receiver Algorithms and Complexity

This section discusses the different receive algorithms and their associated complexity for the reception and decoding of the WUS.

The complex base-band received signal vector $\boldsymbol{y}_{l,p} \in \mathbb{C}^K$ for K sub-carriers, on receive antenna $p = 1,2, \ldots, P$ and OFDM symbol l can be expressed as

$$\boldsymbol{y}_{l,p} = \boldsymbol{H}_{l.p}\boldsymbol{x}_{l,m} + \boldsymbol{n}_{l,p} \text{ ,}$$

where $\boldsymbol{H}_{l.p} = diag(\boldsymbol{h}_{l,p})$ with $\boldsymbol{h}_{l,p} \in \mathbb{C}^K$ the vector of complex channel responses on sub-carriers $k = 1,2, \ldots, K$, $\boldsymbol{x}_{l,m} \in \mathbb{C}^K$ is the transmit vector for message m and $\boldsymbol{n}_{l,p}$ the noise vector.

4.1 Non-Coherent Detection

In the absence of channel state information at the receiver, a near-optimal receive algorithm for the estimate $\hat{m}$ of message m is given by

$$\hat{m} = \arg\max_{m} \sum_{p=1}^{P} \left\| \sum_{l=1}^{L} \boldsymbol{y}_{l,p}^H \boldsymbol{x}_{l,m} \right\|^2$$

Essentially, we compute the correlation of the received signal with all $M = 2^B$ possible transmit signals and choose the message which maximizes the power of the correlation. In terms of complex multiplications (MUL), the non-coherent receiver requires $K_{NZP}PLM$ MUL, where K_{NZP} are the sub-carriers in the transmit signal with *non-zero* power.

4.2 Non-Coherent Detection with Reduced Complexity

With the proposed product code, where the messages in frequency and time-domain can be decoded independently, the NCD can be simplified to reduce the complexity. First the estimate $\widehat{m}_0$ of message m_0 corresponding to the B_0 bits encoded in frequency-domain are computed as

$$\widehat{m}_0 = \arg\max_{m_0} \sum_{p=1}^{P} \sum_{l=1}^{L} \left\| \boldsymbol{y}_{l,p}^H \boldsymbol{f}_{l.m_0} \right\|^2$$

where $\boldsymbol{f}_{l.m_0}$ is the frequency-domain code sequence for m_0 in OFDM symbol l, i.e. the lth column of $\boldsymbol{F}_{m_0}$. This estimation requires $K_{NZP}PLM_0$ MUL.

With the estimate $\widehat{m}_0$ the message m_1 corresponding to the remaining B_1 bits can be estimated according to

$$\widehat{m}_1 = \arg\max_{m_1} \sum_{p=1}^{P} \left\| \sum_{l=1}^{L} \boldsymbol{y}_{l,p}^H \boldsymbol{r}_{l,\widehat{m}_0+m_1M_0} \right\|^2$$

where $\boldsymbol{r}_{l,\widehat{m}_0+m_1M_0}$ is the lth column of $\boldsymbol{R}_{l,\widehat{m}_0+m_1M_0} = \boldsymbol{F}_{\widehat{m}_0} diag(\boldsymbol{w}_{m_1})$, i.e. the transmit signal for message m_1 conditioned on the estimate $\widehat{m}_0$. This detection requires only LPM_1 MUL since $\boldsymbol{y}_{l,p}^H \boldsymbol{r}_{l,\widehat{m}_0+m_1M_0} = \boldsymbol{y}_{l,p}^H \boldsymbol{f}_{l.\widehat{m}_0} w_{l,m_1}$and the term $\boldsymbol{y}_{l,p}^H \boldsymbol{f}_{l.\widehat{m}_0}$ has already been computed.

The performance the low-complexity receiver can be improved by creating a list of several hypothesis for $\widehat{m}_0$. In [1], we show that with a single hypothesis, i.e. $N_{\widehat{m}_0} = 1$, the reduced complexity NCD performs within a fraction of a dB compared to the full NCD and with $N_{\widehat{m}_0} = 2$, the performance is identical. Table 1 provides complexity comparison in terms of complex multiplications.

	Reduced Complexity NCD (RC-NCD) $LPK_{NZP}M_0 + N_{\widehat{m}_0}LPM_1$								**NCD** $K_{NZP}LPM$
$N_{\widehat{m}_0}$	1	2	3	4	5	6	7	8	
$K_{NZP} = 1$	7392	14560	21728	28896	36064	43232	50400	57568	57344
$K_{NZP} = 3$	7840	15008	22176	29344	36512	43680	50848	58016	172032
$K_{NZP} = 6$	8512	15680	22848	30016	37184	44352	51520	58688	344064
$K_{NZP} = 12$	9856	17024	24192	31360	38528	45696	52864	60032	688128

Table 1: NCD complexity comparison in terms of complex multiplications, $L = 14$, $P = 2$, $B = 11$, $B_0 = 3$ $(M_0 = 8)$, $B_1 = 8$ $(M_1 = 256)$.

It can be observed, that the RC-NCD features a significantly reduced complexity compared to the full NCD, especially if the number of non-zero sub-carriers of the frequency-domain sequences increases.

Observation 1: The proposed transmission scheme has low complexity because detection in time and frequency domain can be efficiently separated.

5 Information other than Wake-up

Both PEI and LP-WUS deliver up to 8 and 5 bits, respectively, to allow for group-based wake-up. Since an OFDM-based WUS design is more spectrally efficient, a larger payload (> 8 bits) would be beneficial e.g. to enhance the wake-up functionality by allowing for enhanced grouping or UE-specific wake-up. Even small amounts of data or (wake-up related) control signaling may be considered to further improve potential power-saving gains.

Proposal 1: Study the benefits of increased payload capacity for 6G WUS.

6 Simulation Results

In this section, we evaluate the BLER performance of different sequence-based WUS designs. We assume the same WUS BW as in NR Rel-19, i.e. 11 PRB. The simulation assumptions are summarized in Table 2.

Our evaluation assumes a larger payload of $B = 15$ information bits, which could for instance indicate the wake-up for a specific UE. Encoding 15 bits with a sequence-based design is challenging because of the limited amount of resources, e.g. 132*14=1848 REs, as well as the higher receiver complexity of correlating with 2^{15} sequences.

Thus, we evaluate the Rel-19 WUS sequence design assuming there is *no* Manchester coding and only a single OOK symbol per OFDM symbol (i.e. $M = 1$) with $N_{seq} = 32$ sequences. Hence, 5 bits are transmitted per OFDM symbol and all 15 bits can be decoded within 3 OFDMs symbols. The transmission is then repeated on the remaining OFDM symbols configured for the transmission. The receiver performs OFDM demodulation and correlates with all possible sequences per OFDM symbol. The obtained correlation values are then combined with the repetitions before choosing the highest correlation peak to decode the payload.

The proposed VHC scheme uses the low-complexity receiver with 2 hypotheses for the frequency-domain sequences (VHC-LC2). The BW of 11 PRBs allows for the transmission of $B_0 = 7$ bits with orthogonal sequences and $B_1 = 8$ bits are transmitted over the time-domain. In this simulation, we choose to encode B_0 with different frequency offsets of a single non-zero power RE.

From the results in Figure 2, we observe a 4dB performance gain @1%BLER of the proposed VHC-based WUS design compared to the Rel-19 solution. The B_0 bits offer an additional 1dB gain. This gain results from the fact that the B_0 bits are encoded in frequency domain and repeated over the entire transmission duration. The encoding of the B_1 bits with a linear code designed for non-coherent detection further contributes to the strong overall performance.

We conclude, that there are efficient WUS designs available if the number of available resources is too small to encode the payload with a purely sequence-based approach.

Observation 2: The proposed WUS transmission scheme significantly outperforms the WUS Rel-19 design.

Proposal 2: Study sequence-based WUS designs that allow for increased payload capacity.

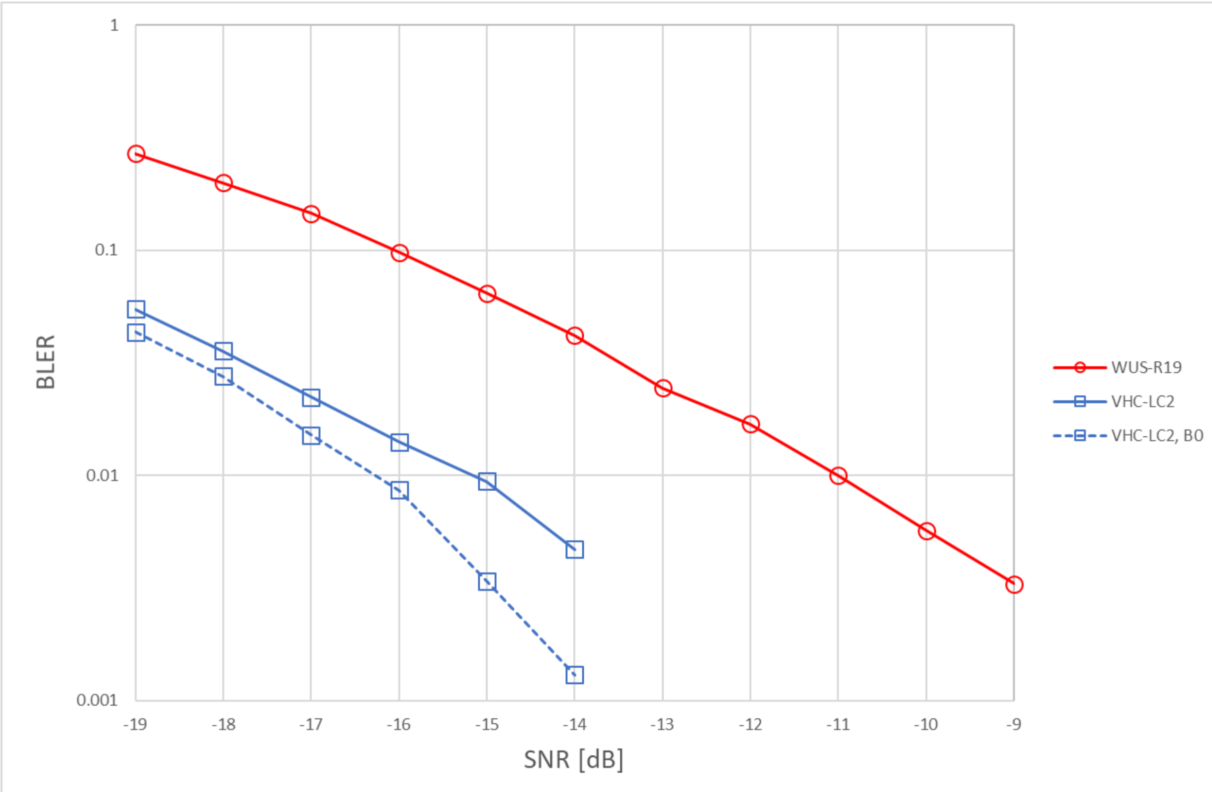

Figure 2: BLER Performance, TDL-C 30ns, $B = 15$ bit, 11 PRBs.

7 Conclusion

In this contribution, the following proposals and observations have been made:

Observation 1: The proposed transmission scheme has low complexity because detection in time and frequency domain can be efficiently separated.

Proposal 1: Study the benefits of increased payload capacity for 6G WUS.

Observation 2: The proposed WUS transmission scheme significantly outperforms the WUS Rel-19 design.

Proposal 2: Study sequence-based WUS designs that allow for increased payload capacity.

8 References

[1] RP-212941, “Discussion on DMRS-less PUCCH for UL Coverage Enhancements”, EURECOM, RAN#94e, Dec, 2021.

9 Appendix

Link-Level simulation assumptions are shown in Table 2 below.

Parameter	**Value**
Carrier Frequency	2.6 GHz (FDD)
Channel BW	20MHz (51 PRBs @ 30kHz SCS)
SCS	30 kHz
Channel Model	TDL-C, 30ns Delay Spread, 0 km/h
Number of receive antennas at UE	2
Number of transmit antennas at gNB	1
Payload size	15 bits
Number of OFDM symbols	14
WUS bandwidth	11 PRB

Table 2: Link-level simulation assumptions.