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Title: Discussion on LP-WUS and LP-SS Design
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1. Introduction

In 3GPP wake-up signals (WUS) are specified in both LTE-M/NB-IoT as well as NR-Rel-17 under the name of paging early indication (PEI). The key difference between those WUS and a low-power (LP)-WUS is that the LP-WUS is received with a *low-power* wake-up receiver (WUR) *independent* of the main radio, i.e. the main radio can be turned off.

A SI [1] on low-power (LP) wake-up signal (WUS) and receiver for NR has been carried out in Rel-18 with the report available in [2]. In Rel-19, a WI [3] was agreed including the following objective related to LP-WUS and LP-SS design:

Objectives:

- To specify an LP-WUS design commonly applicable to both IDLE/INACTIVE and CONNECTED modes (RAN1, RAN4)
  - Specify OOK (OOK-1 and/or OOK-4) based LP-WUS with overlaid OFDM sequence(s) over OOK symbol
    - The LP-WUS design shall ensure that for IDLE/INACTIVE operation, the same information is delivered irrespective of LP-WUR type. The OFDM sequence can carry information.
  - At least duty-cycled monitoring of LP-WUS is supported
- For IDLE/INACTIVE modes
  - Specify procedure and configuration of LP-WUS indicating paging monitoring triggered by LP-WUS, including at least configuration, sub-grouping and entry/exit condition for LP-WUS monitoring (RAN2, RAN1, RAN3, RAN4)
  - Specify LP-SS with periodicity with Yms for LP-WUR, for synchronization and/or RRM for serving cell. (RAN1, RAN4)
    - LP-SS is based on OOK-1 and/or OOK-4 waveform with or without overlaid OFDM sequences. Further down selection between with and without overlaid OFDM sequences is to be done within WI.
    - Note: For LP-WUR that can receive existing PSS/SSS, existing PSS/SSS can be used for synchronization and RRM instead of LP-SS.
    - Y will be decided within WI. 320ms is the start point.
  - Specify further RRM relaxation of UE MR for both serving and neighbor cell measurements, and UE serving cell RRM measurement offloaded from MR to LP-WUR, including the necessary conditions (RAN4, RAN2)

This contribution focuses on LP-WUS design in RRC IDLE mode.

2. OOK Waveform Designs

Regarding the LP-WUS waveform design the following has been agreed in RAN#116-bis:

> Agreement:
>
> For OOK-4 with M >1, support M=2 & M=4 (working assumption) for LP-WUS.
>
> - FFS whether value of M depends on SCS
> - FFS M=1 for OOK-4

In this section, we review the OOK modulation and the 2 agreed OOK schemes, OOK-1 and OOK-4.

We denote N PRBs as the bandwidth of the LP-WUS including potential guard bands (GB). The actual number of used PRBs for the WUS and GB are referred to as N_{WUS} and N_{GB}, respectively, such that $N = N_{WUS} + N_{GB}$.

Moreover, we refer to the ON-signal $\boldsymbol{a}$ as the sequence of complex symbols allocated to the OOK symbol if the corresponding bit is one.

The overall DL transmission block diagram is depicted in Figure 1. The B information bits $\boldsymbol{b} = [b_0, b_1, \dots, b_{B-1}]$ are encoded and the resulting C coded bits $\boldsymbol{c} = [c_0, c_1, \dots, c_{C-1}]$ are modulated onto L consecutive OFDM symbols each carrying M bits. That is, M is the number of coded bits per OFDM symbol or the number of OOK symbols. Subsequently, the signal in frequency-domain $\boldsymbol{S}_m$ of message $m = 0,1, \dots, 2^B - 1$ is mapped to the overall resources $\boldsymbol{X}_m$ of N sub-carriers and OFDM-modulated resulting in the time-domain signal $\boldsymbol{x}_m(t)$.

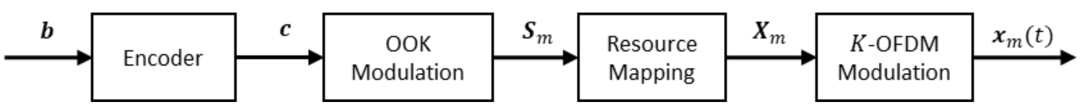

Figure 1: Overall transmission block-diagram per OFDM symbol.

Subsequently, we review OOK modulation as well as the specific OOK-1 and OOK-4 modulations in the WI objectives.

2.1. OOK Modulation

Denote the OOK modulated signal $\boldsymbol{s}_m$ in time-domain for message $m = 0,1, \dots, 2^C - 1$ as

$$\boldsymbol{s}_m = [\boldsymbol{s}_0, \boldsymbol{s}_1, \dots, \boldsymbol{s}_{C-1}]$$

where C is the number of coded bits transmitted per OFDM symbol l and $\boldsymbol{s}_i = [s_{i0}, s_{i1}, \dots, s_{iN'_B-1}]$ with $N'_B = \lfloor N'/C \rfloor$ the number of samples for sequence $\boldsymbol{s}_i$, $i = 0,1, \dots, C-1$. The OOK modulation consists of mapping an ON-signal or ON-sequence $\boldsymbol{a} = [a_0, a_1, \dots, a_{N'_B-1}]$ to $\boldsymbol{s}_i$ whenever the corresponding bit is one, i.e.

$$s_i = \begin{cases} \mathbf{0}, & c_i = 0 \\ \boldsymbol{a}, & c_i = 1 \end{cases}$$

Hereby, the ON-signal $\boldsymbol{a}$ can be *any complex sequence* and can be different for every $\boldsymbol{s}_i$.

In case of multiple ON-sequence, consider Q ON-sequences $\boldsymbol{a}_q$ able to encode $\mathrm{B}_2 = \log_2 Q$ bits.

2.2. OOK-1: Single-bit in 1 OFDM symbol

The simplest OOK scheme allocates the ON-signal $\boldsymbol{a}$ of length N_{WUS} to the corresponding WUS SCs if the bit is one and zeros otherwise (in baseband).

A block-diagram is shown in Figure 2.

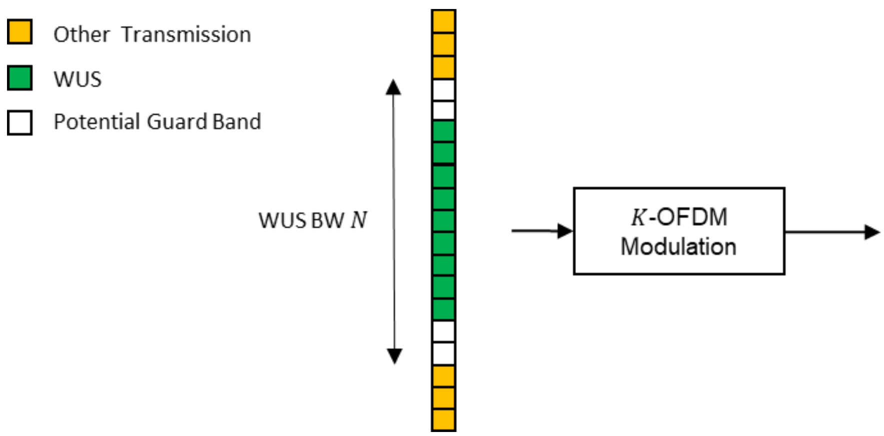

Figure 2: Block-diagram for OOK-1.

This transmission scheme can only transmit a single bit per OFDM symbol. Thus, the only means to increase the data rate is to shorten the OFDM symbol length by increasing the sub-carrier spacing (SCS).

2.3. OOK-4: Transform M-bit OOK in time domain

This scheme uses DFT-precoding to convert M OOK symbols in time-domain to frequency domain for allocation to N_{WUS} PRBs.

An example block-diagram is depicted in Figure 3. From the M (coded or physical) bits, a signal of length N' is generated where each bit is mapped to a sequence $\boldsymbol{s}_i$ of length N'_B, the sequence is 0 if the corresponding bit is zero and $\boldsymbol{s}_i = \boldsymbol{a}$ otherwise. The resulting sequence is DFT precoded and the output is mapped to the overall transmission bandwidth.

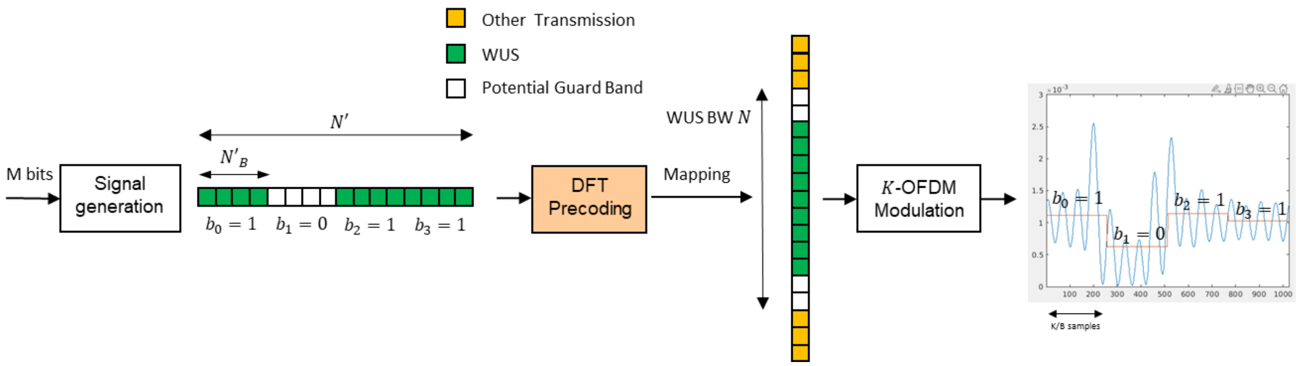

Figure 3: OOK-4, Block-diagram with M=4.

2.4. Waveform Generation

For the discussion on LP-WUS waveform generation, we borrow the block-diagram from the feature lead summary in Figure 4.

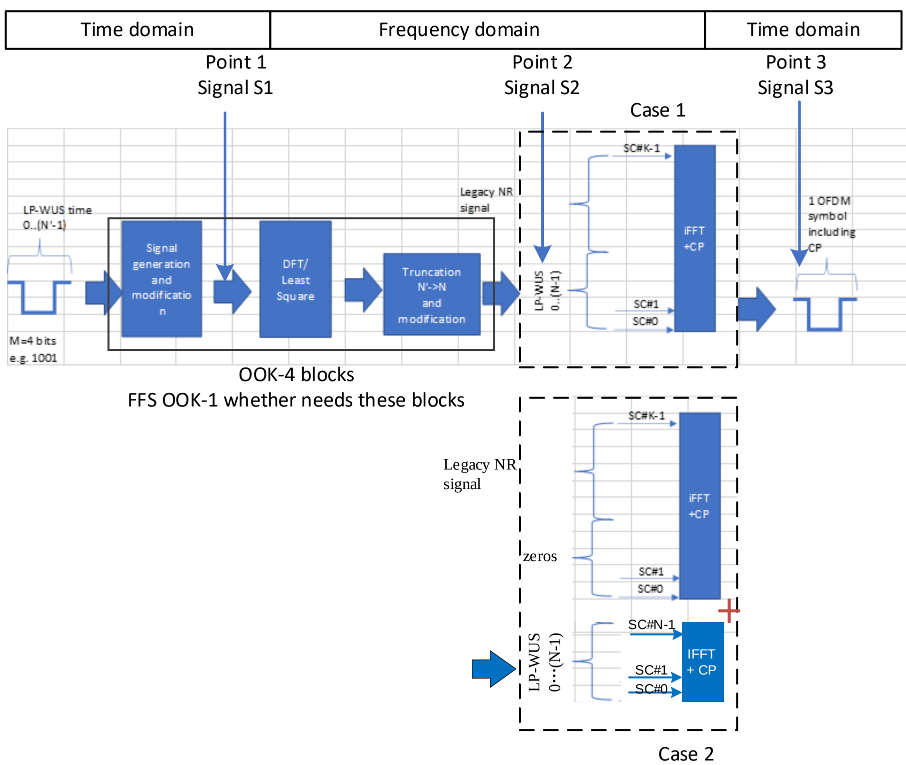

Figure 4: Block-diagram for LP-WUS waveform generation.

2.4.1. Pulse Shaping at S1

Pulse shaping should be considered together with the ON-sequence design. It has been shown that shorter pulses, or ON-sequences that concentrate the energy in the middle of the pulse, are more robust to timing errors. However, short pulses will capture less multi-path diversity and it is therefore necessary to ensure that the duration of the pulse is sufficiently long.

Moreover, a potential preamble and/or a receiver that applies time-windowing (sliding window), will mitigate the impact of timing inaccuracy. Hence, shortening the pulse might not be necessary/beneficial after all.

Proposal 1: Consider if pulse-shaping is required after sequence design and potential preamble are agreed.

2.4.2. DFT-Shift for Signal at S1

Given the generation of the OFDM waveform, a DFT-shift might be required to ensure that the LP-WUS spectrum integrates seamlessly into the overall NR spectrum.

For an energy detector, the DFT-shift has no impact on the detection performance. However, a coherent receiver needs to know if a DFT-shift has been applied or not. In general, the DFT shift can be applied at the gNB or at the receiver. We don't have a strong preference but since the standard MR OFDM-based receiver compensates for shift, the same could be assumed for the LR.

Proposal 2: The DFT-shift is compensated at the LR.

2.4.3. Mapping of WUS at S2 to existing NR signals

It has been proposed to map the frequency-domain WUS to existing NR signals in order to reduce gNB complexity and to improve robustness to frequency errors. In general, quantizing or mapping, the frequency-domain WUS will change its properties in time-domain. From a standard perspective, such a mapping is not necessary since arbitrary values are allowed in frequency domain, e.g. consider precoded signals for MU-MIMO which surely are different from QAM constellation or existing sequences.

Moreover, for OOK-4 such mapping is difficult and needs to ensure that the performance of the energy detection receiver is not impaired. Hence all supporters should evaluate the performance of both OFDM sequence detection *and* energy detection.

Proposal 3: Do not consider mapping/quantizing WUS in frequency-domain.

3. Coding

The WUS coding procedure is shown in Figure 5. The payload $\boldsymbol{b}$ is channel encoded resulting in codeword $\overline{\boldsymbol{d}}$ which is then rate matched to the available resources. Subsequently, the rate matched coded bits $\boldsymbol{d}$ are encoded with a line code, e.g. Manchester coding or Pulse Position coding, resulting in the coded bits sequence $\boldsymbol{c}$ which is OOK modulated to a time-domain sequence $\boldsymbol{s}(t)$. This time domain sequence is converted to frequency-domain and mapped to the WUS bandwidth.

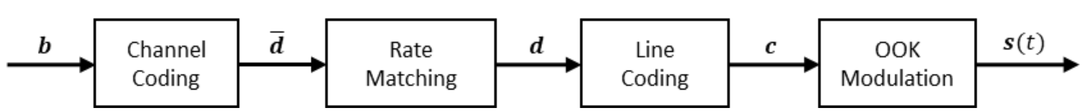

Figure 5: Block-diagram of WUS encoding procedure.

3.1. Channel Coding and Rate Matching

To meet target requirements, it has been agreed to consider the following scheme:

Agreement:	RAN1#120

For LP-WUS information carried by OOK, select alt 1 for codepoint to meet FAR and MDR performance.

- Alt 1: Coding with rate matching if any
  - For more than 2 information bits: RM coding as in section 5.3.3.3 of 38.212 (FFS: whether scrambling in the same section is applied or not)
  - 2 information bits (if supported): coding as in section 5.3.3.2 of 38.212 (FFS: whether scrambling in the same section is applied or not)
  - 1 information bit (if supported): repetition as in section 5.3.3.1 of 38.212 (FFS: whether scrambling in the same section is applied or not)
  - FFS: Code block length
  - FFS: Further repetition before or after Manchester coding

The input bit sequence $\boldsymbol{b}$ to the channel coder consists of at most 5 bits for RRC_IDLE/INACTIVE mode and at most 8 or 16 bits for RRC_CONNECTED mode.

For both RRC_IDLE/INACTIVE and RRC_CONNECTED (working assumption), a codepoint-based mapping has been agreed, where the UE has to monitor up to 2 or 8 WUS per MO, respectively. A codepoint is a sequence of bits triggering the wake-up of one or multiple groups. Hence, a codepoint defines a mapping between subgroups and the input bits sequence $\boldsymbol{b}$. An example is shown in Table 1 for 16 subgroups and a $B = 3$ bit payload.

Subgroups	Mapping to payload/codepoint $\boldsymbol{b}$
1	$\{0,0,0\}$
2,3	$\{0,0,1\}$
4	$\{0,1,0\}$
5,6,7	$\{0,1,1\}$
8,9	$\{1,0,0\}$
10,11,12,13	$\{1,0,1\}$
14,15,16	$\{1,1,0\}$
1-15	$\{1,1,1\}$

Table 1: Example of subgroup to codepoint/payload mapping for 16 subgroups and payload of B=3 bits.

Thus, to our understanding, a codepoint $\boldsymbol{b}$ is defined before channel coding.

The codepoint is subsequently channel coded and the output is rate matched to the available/configured WUS resources. RM coding is used if $B > 2$, however RM coding is only defined for up to $B = 11$. The question is then how payloads of $B > 11$ bits, e.g. in RRC_CONNECTED, are encoded?

Observation 1: Consider a channel encoding procedure for more than 11 information bits, e.g. 16 bits in RRC_CONNECTED mode.

After rate matching, the coded bits $\boldsymbol{d}$ are line coded with code rate $R = 1/2$.

Subsequently, we propose a line coding scheme other than Manchester Coding called Pulse Position Coding for $M = 4$. Contrary to the other alternatives, PPC has *no additional resource overhead* but achieves a SNR gain of 3dB simply by pooling the available power to a single OOK symbol.

3.2. Line Coding: Discussion on Manchester Coding

The study on LP-WUS [2] suggests to utilize Manchester Coding, where one input bit is mapped to two coded bits. This coding technique is especially adapted to OOK with ED since it does not require threshold detection but a simple energy comparison of the two OOK symbols is sufficient for reliable decoding.

The following agreement has been made

Agreement: Support Manchester coding for LP-WUS • FFS other coding schemes.	**RAN1#118**

To further refine this agreement, we suggest to agree on how to specify the mapping:

1: 0 →[0 1] and 1→[1 0] or

2: 0 →[1 0] and 1→[0 1]

As discussed in Ambient IoT, the IEEE 802.3 convention is Option 2 because the encoded signal will just be an XOR of the clock and the data. As a result, the following agreement has been reached in A-IoT:

Agreement
The study assumes the following bit to chip mapping for Manchester encoding:
  - bit 0→chips{10}, bit 1→chips{01}
- FFS: Variant of the above for CP handling

Therefore, we suggest to follow this MC mapping.

Proposal 4: Specify Manchester Coding as 0 →[1 0] and 1→[0 1].

Subsequently, we address the FFS point on other coding schemes in the above agreement.

In case of $M = 4$, following the Manchester coding scheme, two of the four available OOK symbols per OFDM symbol will be ON. Hence, the available transmit power per OFDM symbol is divided among the two ON-symbols.

In the next section we discuss an alternative approach where two input bits are encoded in the position of a single ON symbol.

3.3. Line Coding: Pulse Position Coding for $M = 4$

The detection performance of Manchester Coding can be improved by jointly encoding multiple bits. More precisely, B input bits $\boldsymbol{b} = b_0, b_1, \dots, b_{B-1}$ are encoded to C output bits $\boldsymbol{c} = c_0, c_1, \dots, c_{C-1}$ according to

$$\bar{\boldsymbol{c}} = 2^m$$

where $m = 0,1, \dots, 2^B - 1$ is the message and $\bar{\boldsymbol{c}}$ is the *decimal* representation of codeword $\boldsymbol{c}$. For $B = 1$, we obtain the conventional Manchester code of rate $R = 1/2$ presented in Table 2.

Input Bits	Coded Bits
0	01
1	10

Table 2: Encoding for B=1 input bits and C=2 coded bits, rate R = ½.

Encoding $B = 2$ bits into $C = 4$ coded bits is given by Table 3.

Input Bits	Coded Bits
00	0001
01	0010
10	0100
11	1000

Table 3: Encoding for B=2 input bits and C=4 coded bits, rate R = ½.

As can be seen from Table 3, the input bits are encoded in the position of the one-bit, hence the name Pulse Position Coding (PPC), or Pulse Position Modulation.

Figure 6 shows an example of the transmitted waveform for the two encoding schemes without any co-scheduled transmissions. It can be observed that the pulse position coded data results in higher signal amplitudes, i.e. a power boost, compared to Manchester coding.

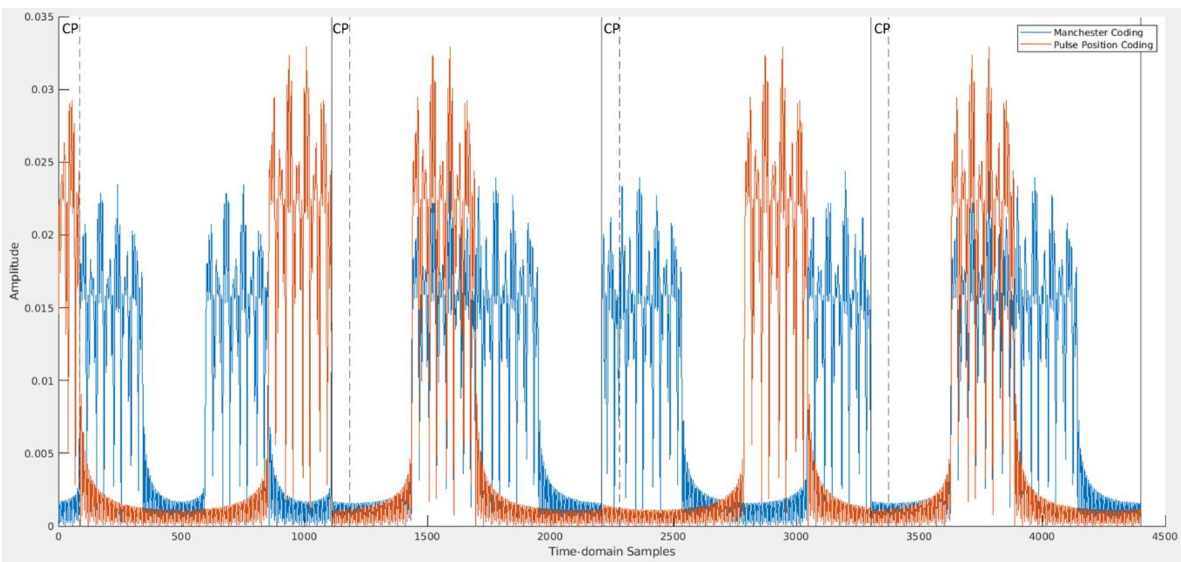

Figure 6: Transmit Signal for payload $\boldsymbol{b} = [1\ 1\ 0\ 1\ 1\ 0\ 0\ 1]$, $M = 4$, 4 OFDM symbols, no ACI.

The main advantage of using $B = 2$ compared to $B = 1$, is that the total WUS transmit power per OFDM symbol is concentrated into a single ON-symbol and not divided among two ON-symbols. Hence, there is a 3dB SNR gain.

From results of the SI [2], it has been observed that the case $M = 4$ has the worst coverage compared to $M = 2$ and $M = 1$ with a loss of about 3dB for the reasons explained above. Thus, any coding scheme that can mitigate the coverage loss should be considered.

Observation 2: $\boldsymbol{M = 4}$ with Manchester Coding has the worst coverage compared to $\boldsymbol{M = 1, 2}$.

The proposed PPC results in a 3dB SNR gain compared to Manchester Coding and hence the required SNR is the same as for $M = 2$ but with double the data rate. Therefore, the WUS payload can be delivered twice as fast reducing the power consumption of the LR.

As eluded to in Section **Erreur ! Source du renvoi introuvable.**, PPC and mapping to OOK symbols can be viewed as OOK sequence selection, where the bits are jointly encoded into *orthogonal* sequences.

Proposal 5: For $M = 4$, consider jointly encoding multiple bits into ON pulse position to increase SNR by 3dB.

3.3.1. Decoding at the Receiver

At the receiver, one can distinguish two ways to decode a Manchester encoded payload, (i) per-codeword decoding and (ii) sequence decoding.

In case of per-codeword decoding, the Manchester codeword is decoded by accumulating the energy of the two corresponding OOK symbols and comparing their energy. The same strategy applies to PPC where the energy is computed for four OOK symbols and the symbol with the maximum energy determines the information bits.

For sequence-decoding, multiple codewords are decoded jointly by correlating with all possible coded sequences. Naturally, this is only feasible if the transmitted payload is small, otherwise the number of possible transmitted messages and hence the number of correlations increases exponentially. An important factor that needs to be accounted for is interference, which changes over time and can be significantly different at the start of the received sequence compared to the end of the sequence. This issue is usually addressed by taking the difference between the amplitudes of two adjacent OOK symbols which constitute a codeword under the assumption that the interference remains approximately constant within the codeword.
In case of PPC, the same paradigm applies but instead of subtracting the amplitude of the other OOK symbol, one has to subtract the *average* of the other *three* amplitudes because the codeword contains 4 OOK symbols. Again, the assumption here is that the interference is approximately constant over the codeword which, compared to the case of Manchester coding, is double the duration.

Discussion of Self-Clocking Aspect

Manchester coding has the advantage the there is always a transition from zero to one or one to zero in the codeword which allows for timing alignment, also referred to as self-clocking. One simple implementation to maintain timing alignment during the decoding process is an Early-Late architecture, where two parallel branches accumulate energy with one branch starting accumulation a few samples earlier (Early branch) and the other a few samples later (Late branch). By comparing the results of the two branches, one can determine if it is necessary to realign the OOK symbol timing.

With PPC, there is only one transition in the codeword but the same principle applies. The only difference is that the timing alignment is evaluated at the end of the codeword which takes four OOK symbols instead of two. However, unless the timing changes significantly within 4 OOK symbols, there is no impact on decoding performance.

Also note that timing determination, i.e. determining the rising or falling edges of the pulse, is improved in PPC since the SNR of the Pulse is higher.

3.3.2. Discussion on PAPR

It has been pointed out that PPC increases the PAPR of the resulting OFDM time-domain waveform which impacts existing base-station implementations, e.g. they may fail RAN4 requirements.

In this section, we analyze the impact of PPC on Peak-to-Average Power Ratio (PAPR) of the transmitted waveform. Unless stated otherwise, the simulation assumption in Table 9 are used. The transmission power of the WUS per OFDM symbol is kept constant. For an ON-sequence $\boldsymbol{a} = [a_0 a_1 \dots a_{N'_B-1}]$ of length N'_B, a *truncated* Zadoff-Chu sequence is used which is generate as

$$a_n = e^{\frac{-j\pi u n(n+1)}{N_{ZC}}}$$

where $n = 0,1,\dots,N'_B - 1$ and $N_{ZC} = NP(N'_B)$ with $NP(x)$ denoting the next prime greater or equal to x. In the simulations we use root $u = 1$.

From the encoding, it is expected that the PAPR of PPC is 3dB higher than with MC. However, the increase in PAPR is lower, for instance in Figure 6, the PAPR is 6.8dB and 9.3dB for MC and PPC, respectively. Hence the increase is only 2.5dB because of the CP. More precisely, if the last OOK pulse within the OFDM symbol is ON, the last samples of that symbol will be copied to the CP which increases the mean power of the signal and thus decreases the PAPR. The reduction in PAPR due to the CP is especially large in PPC because of the larger power in the ON symbol which results in a much lower PAPR for those payloads, overall resulting in a reduced average PAPR

With co-scheduled 64-QAM transmissions, the transmit signal looks quite different as depicted in Figure 7. The PAPR increases for MC and PPC to 9.3dB and 11.2dB, respectively.

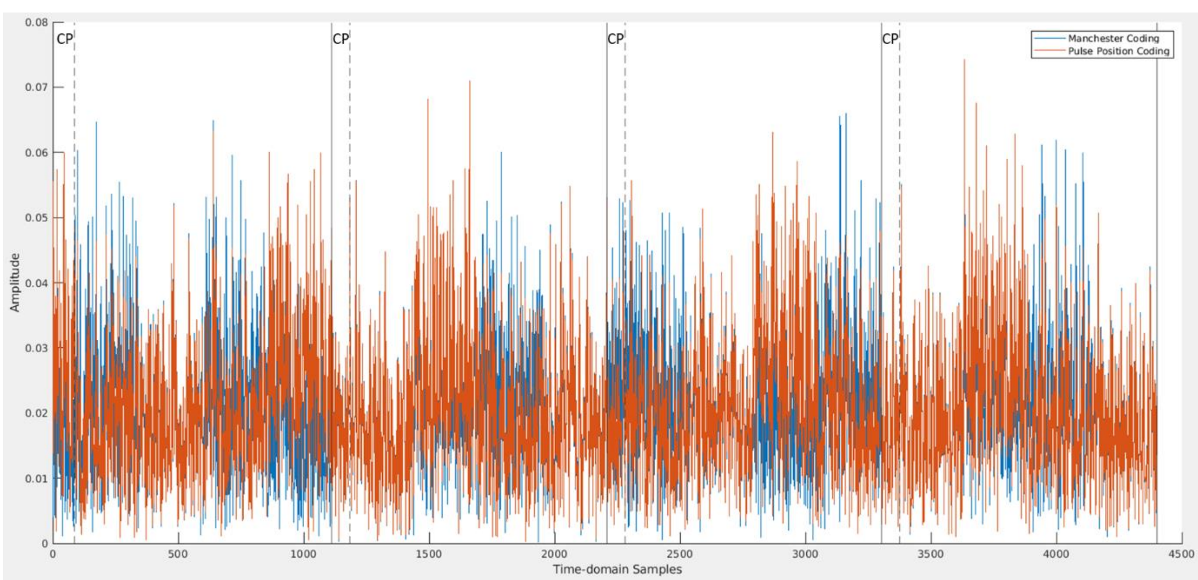

Figure 7: Transmit Signal for payload $\boldsymbol{b} = [11011001]$, M=4, 4 OFDM symbols, with 64-QAM ACI.

Figure 8 shows the CCDF of the PAPR for both coding schemes in the absence of ACI, i.e. LP-WUS is the only transmission in the 20 MHz BW. It can be observed that the mean PAPR is 6.74 dB and 9.37 dB for MC and PPC, respectively, i.e. PPC increase the mean PAPR by about 2.63 dB.

In both schemes, the difference between average PAPR and 1% outage PAPR is about 0.3dB. This means that only 1% of the time, the PAPR exceeds the average PAPR by 0.3dB.

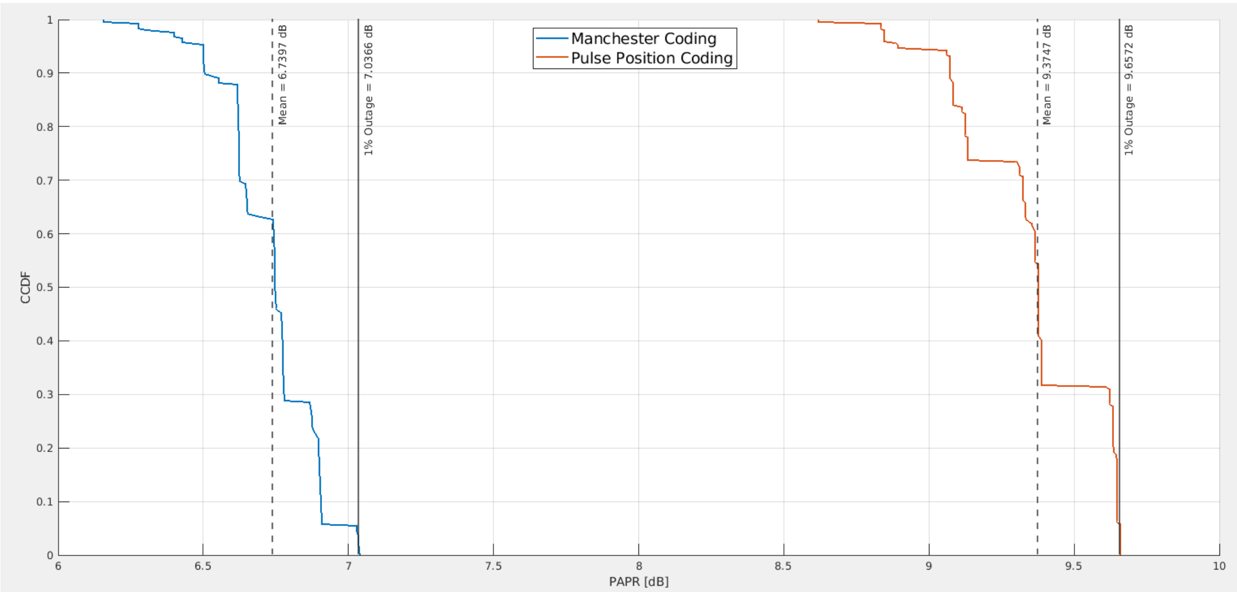

Figure 8: CCDF of PAPR, no ACI, 20MHz, 30kHz SCS, M=4, 8 bit payload, 4 OFDM symbols, average over 10k realizations.

Figure 9 shows the CCDF *with* co-scheduled 64-QAM transmissions. The mean PAPR is 9.52dB and 10dB for MC and PPC, respectively, i.e. a difference of about 0.5dB. As expected, the 64-QAM transmissions increase the average PAPR as well as the 1% outage PAPR. The transmission exceeds the average PAPR by about 1.6 dB for both schemes, 1% of the time.

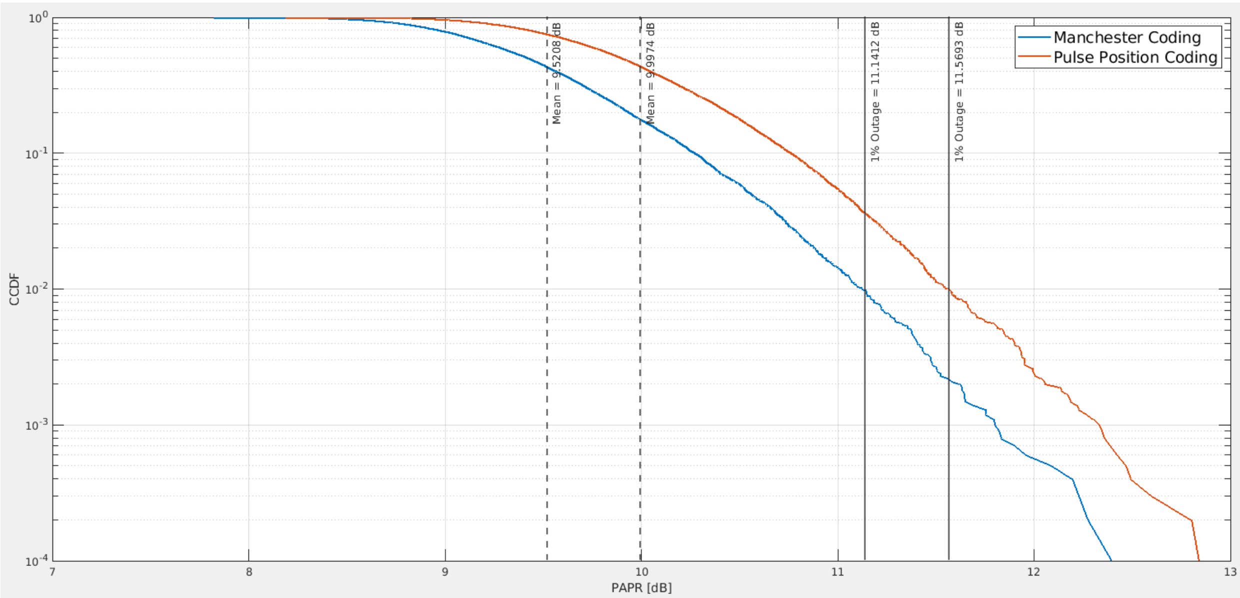

Figure 9: CCDF of PAPR, 64-QAM ACI, 1PRB GB, 20MHz, 30kHz SCS, M=4, 8 bit payload, 4 OFDM symbols, average over 10k realizations.

In summary, the impact of PPC on PAPR is manageable by the transmitter. With co-scheduled transmissions the average and 1% outage PAPR are only 0.5dB larger with PPC than MC. Without co-scheduled transmissions, the PAPR is lower for MC but still well within the acceptable range for PPC.

Table 4 provides simulation results for PAPR of various transmission schemes and system configurations.

Scheme	**20MHz, SCS 30, 51PRBs, FFT=1024**	**20MHz, SCS 15, 106 PRBs, FFT=2048**	**10MHz, SCS 15, 52 PRBs, FFT=1024**	**100MHz, SCS 30, 273 PRBs, FFT=4096**	**20MHz, SCS 15, 106 PRBs, FFT=2048, WUS BW=2.5MHz**
64QAM	9.5/11.0	9.8/11.3	9.5/11.1	10.1/11.5	9.8/11.3
OOK-1, R=1/2	9.6/11.3	9.9/11.3	9.4/11.0	10.1/11.6	9.8/11.4
OOK-4, M=1, R=1/2	9.5/11.1	9.5/11.1	9.5/11.0	10.1/11.5	9.8/11.4
OOK-4, M=2, R=1/2	9.7/11.4	9.9/11.3	9.5/11.0	10.1/11.5	9.9/11.4
OOK-4, M=4, R=1/2	9.5/11.1	10.0/11.6	9.9/11.5	10.1/11.5	9.8/11.4
OOK-4, M=4, PPC R=1/2	10.0/11.6	10.7/12.2	11.3/12.7	10.2/11.6	10/11.6
OOK-4, M=4, PPC R=1/2, 64-QAM quantization	10.0/11.6	10.6/12.2	11.2/12.7	10.2/11.6	10/11.6

Table 4: Average PAPR [dB] / 1%Outage PAPR [dB], Simulation results for different encoding schemes and system configurations. WUS BW either 11 PRBs or 22 PRBs

The baseline "64QAM" assumes random 64-QAM symbols on all sub-carriers, i.e. no WUS. It can be observed that the impact on PAPR increases with larger WUS BW w.r.t. to the system BW. Moreover, up to $M = 2$, using OOK-1 or OOK-4 does *not* increase PAPR w.r.t. to the baseline. For $M = 4$ with Manchester coding R= $1/2$, there is only a very small increase of PAPR (0.4 dB) compared to the baseline.

For $M = 4$, the PAPR increase caused by PPC is about 0.5 dB and 0.7dB compared to MC if the WUS BW constitutes about 25% of the overall BW (5MHz out of 20MHz) for SCS 30kHz and 15kHz, respectively. For a 10MHz channel and 5MHz WUS the PAPR increase is about 1.4 dB. On the other hand, for a 100MHz channel and 5MHz WUS there is no increase in PAPR. Similarly, for a 20MHz channel with a 2.5MHz WUS the PAPR increase is only 0.2dB. In addition, quantizing the frequency-domain signal to 64-QAM constellation points does not impact the PAPR.

Observation 3: PAPR increase of Pulse Position Coding for $M = 4$ compared to Manchester encoding depends on the ratio of channel BW to WUS BW and is minor (~0.1dB) for many system configurations.

Figure 10 shows the performance for $M = 4$ (and for comparison $M = 2$ with 4 bit payload) with Manchester coding (MC) and Pulse Position Coding (PPC) for a 20MHz BW channel and a WUS BW of 5.04 MHz (24 PRB WUS + 2 PRB GB on each side), solid lines, and 2.52 MHz (12 PRB WUS + 1 PRB GB on each side), dashed lines. It can be observed that there is a 3dB SNR gain of PPC vs. MC in both system configurations. Moreover, in this setting (depending on RX filter design etc.), reducing the WUS BW by half, results in an SNR loss of about 3dB. The performance of MC with $M = 2$ is about the same as $M = 4$ with PPC but only delivering *half* the bit rate.

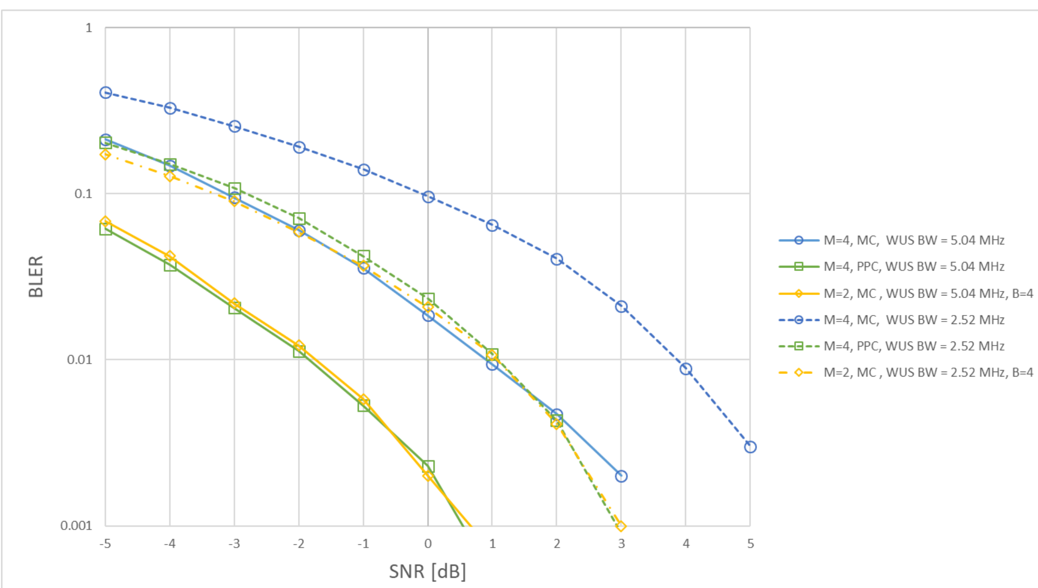

Figure 10: BLER vs. SNR, ED-WUR, 20 MHz (106 PRBs), SCS=15 kHz, M=4, 8 bit payload.

Therefore, with the *same* performance, a network operator can *save* 2.52 MHz of BW by configuring a WUS BW of 2.52 MHz with PPC as opposed to a WUS BW of 5.04 MHz with Manchester coding. The difference in PAPR is 0dB (10dB in both configurations) which is only 0.2dB above the baseline of 9.8dB.

Alternatively, the operator can use $M = 2$ but with half the bit rate.

In summary, we propose PPC for $M = 4$ to pool power in a single OOK symbol resulting in a 3dB gain. PPC is a simple line coding scheme. Other ways to achieve a 3dB gain is to increase the sequence length by more than factor 2 or by repetition. Both alternatives more than double the amount of transmission resources! Hence PPC is a real advantage and should be considered.

Proposal 6: Allow configuration of *Pulse Position Coding* for $\boldsymbol{M = 4}$.

4. Overlaid OFDM Sequences

In this section, we discuss the aspects related to the sequence(s) utilized on the OOK ON-symbols.

4.1. Overlaid OFDM Sequence Design

Concerning the specification of the overlaid OFDM sequence, it has been agreed to support Option 1-1 for OOK-4 $M > 1$:

> Agreement: **RAN1#118**
>
> For overlaid OFDM sequences for LP-WUS, support option 1-1 for OOK-4 M>1.
>
> - Option 1-1: overlaid sequence(s) are the sequence(s) of an OOK on symbol before DFT/LS processing

In RAN1#120 is has been agreed to use a *cyclically extended* Zadoff-Chu sequence $s'(n)$ of length L_{ZC} with cyclic shift C_v.

> Agreement: **RAN1#120**
>
> For M=2, 4, B_{ZC} is given by the largest prime number such that $B_{ZC} < L_{ZC}$, L_{ZC} is the overlaid OFDM sequence length.
>
> - The base overlaid sequence $\mathrm{s(n)}$ is generated by extension of $\boldsymbol{X_q}$
>
> $$\mathrm{s(n)} = X_q\left(n \bmod B_{ZC}\right),\ n = 0,\ldots,L_{ZC}-1$$
>
> - With CS(s) applied to the base overlaid OFDM sequence if any: C_v denotes the potential cyclic shift (s)
>
> $$\mathrm{s'(n)} = X_q\left((n + C_v) \bmod B_{ZC}\right),\ n = 0,\ldots,L_{ZC}-1$$
>
> - Note it doesn't preclude any pulse shaping scheme if any.

Concerning the length L_{ZC} of the time-domain sequence $s'(n)$ the following agreement has been made

> Agreement: **RAN1#120**
>
> At least for M>1, for the overlaid OFDM sequence length, support
>
> - Alt2: $L_{ZC} = 12 * X/M$
> - Note: X is the number of RBs of LP-WUS/LP-SS bandwidth (blanked guard RBs are not included)

With Alternative 2 and $X = 11$ PRBs, there are 3 different ON-sequence lengths, 132, 66 and 33 samples for $M = 1{,}2$ and 4, respectively. Naturally, a longer sequence length can support a larger number of sequences given similar performance. Therefore, it makes sense that the *configured* number of candidate sequences depends on M, i.e. the sequence length. It follows that specifying a small number of "maximum number of candidate sequences", e.g. 4, limits the potential performance for small M.

Concerning the maximum number of candidate overlaid sequences for LP-WUS, the latest agreements read:

Agreement: **RAN1#119** Regarding the maximum number of candidates overlaid sequences to carry LP-WUS information per OOK ON chip for one cell: • support maximum 4 candidates overlaid sequences for M=4

Agreement: **RAN1#120** For idle mode, regarding the maximum number of candidates overlaid sequences to carry LP-WUS information per OOK ON chip for one cell: • support maximum 16 candidates overlaid sequences for M=1 • support maximum 8 candidates overlaid sequences for M=2 • For candidate overlaid sequences across all OOK ON chips of LP-WUS, the number of roots (in specification) is up to [FFS: X], FFS whether the number of roots can be different for different M value.   ○ FFS: The number of overlaid sequences applicable for a UE is no more than 2 per OOK ON chip.

Concerning the FFS part on the number of roots. The concern raised by companies was that with an increased number of roots, the UE has to carry out more correlations. Indeed, if there is only a single root, only one correlation is necessary as the candidate sequence can be determined via the position of the correlation peak. The UE has to monitor only a limited number of codepoints and hence wake-up sequences. The optimal receiver would correlate the entire WUS with those known sequences as described in Section 6.1.2.2. It follows that the number of roots or maximum number of overlaid sequences per OOK ON chip is irrelevant. On the other hand, if the receiver implements a per OOK chip correlation, cf. Section 6.1.2.1, the number of correlations per OOK chip matters. There is a trade-off between the number of roots and the number of required correlations. Many roots and a hence a reduced number of cyclic shifts will result in an increased robustness against timing errors, since the timing window related to the peak position is larger. But it will also result in more correlations.

What is important is that the time between 2 cyclic shifts is the same across different M so that the robustness to timing errors (of the sequence detection) is roughly the same for different M.

In the following, we propose to use only cyclic shifts to compute the candidate sequences.

If the maximum number of candidate sequences is 4, then there are 4 shifts defined. If the gNB only configures 2 sequences then those 2 sequences are chosen among the 4 candidate sequences in a way to maximize the distance between the cyclic shifts.

Define the maximum number of candidate sequences as Q_{max}, the cyclic shifts are given by

$$C_v = \left\lfloor \frac{L_{ZC}}{Q_{max}} \right\rfloor i$$

where $i = 0,1, \ldots, Q_{max} - 1$. This ensures that a maximum identical cyclic shift among the candidate sequences.

The following table details the cyclic shift values for the 3 sequence lengths and different values of Q_{max}.

Sequence Length L_{ZC}	$\boldsymbol{Q = 2}$	$\boldsymbol{Q_{max} = 4}$	$\boldsymbol{Q_{max} = 8}$	$\boldsymbol{Q_{max} = 16}$	$\boldsymbol{Q_{max} = 32}$	$\boldsymbol{Q_{max} = 64}$
132	{0, 66}	{0, 33, 66, 99}	{0, 16, 32, 48, 64, 80, 96, 112}	{0, 8, 16, 24, 32, 40, 48, 56, 64, 72, 80, 88, 96, 104, 112, 120}	{0, 4, 8, 12, 16, 20, 24, 28, 32, 36, 40, 44, 48, 52, 56, 60, 64, 68, 72, 76, 80, 84, 88, 92, 96, 100, 104, 108, 112, 116, 120, 124}	{0, 2, 4, 6, 8, 10, 12, 14, 16, 18, 20, 22, 24, 26, 28, 30, 32, 34, 36, 38, 40, 42, 44, 46, 48, 50, 52, 54, 56, 58, 60, 62, 64, 66, 68, 70, 72, 74, 76, 78, 80, 82, 84, 86, 88, 90, 92, 94, 96, 98, 100, 102, 104, 106, 108, 110, 112, 114, 116, 118, 120, 122, 124, 126}
66	{0, 33}	{0, 16, 32, 48}	{0, 8, 16, 24, 32, 40, 48, 56}	{0, 4, 8, 12, 16, 20, 24, 28, 32, 36, 40, 44, 48, 52, 56, 60}	{0, 2, 4, 6, 8, 10, 12, 14, 16, 18, 20, 22, 24, 26, 28, 30, 32, 34, 36, 38, 40, 42, 44, 46, 48, 50, 52, 54, 56, 58, 60, 62}	{0, 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11, 12, 13, 14, 15, 16, 17, 18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28, 29, 30, 31, 32, 33, 34, 35, 36, 37, 38, 39, 40, 41, 42, 43, 44, 45, 46, 47, 48, 49, 50, 51, 52, 53, 54, 55, 56, 57, 58, 59, 60, 61, 62, 63}
33	{0, 16}	{0, 8, 16, 24}	~~{0, 4, 8, 12, 16, 20, 24, 28}~~	~~{0, 2, 4, 6, 8, 10, 12, 14, 16, 18, 20, 22, 24, 26, 28, 30}~~	~~{0, 1, 2, 3, 4, 5, 6, 7, 8, 9, 10, 11, 12, 13, 14, 15, 16, 17, 18, 19, 20, 21, 22, 23, 24, 25, 26, 27, 28, 29, 30, 31}~~	N/A

Table 5: Cyclic shift values C_v as a function of sequence length L_{ZC} and maximum number of candidate sequences Q_{max}.

As is obvious from Table 5, the potential amount of sequences that need to be stored at the gNB and UE is large. However, the number can be reduced, be reusing sequences for various number of configured candidate sequences $Q \leq Q_{max}$. For instance, consider $Q_{max} = 16$ for all values of M. For $M = 1$

($L_{ZC} = 132$), there are 16 sequences of length 132. If the gNB configures only $Q = 8$, those sequences are chosen among the set of 16 sequences, e.g. by taking every second cyclic shift value. For $Q = 4$, this works as well but is sub-optimal since, by taking every 4th value, the cyclic shift values would be {0, 32, 64, 96} compared to the optimal {0, 33, 66, 99}. However, the performance impact is likely minor and well justified by the reduction in the number of sequences stored at the gNB/UE.

The root q could be dependent on the physical cell ID N_{ID}^{cell} and the cyclic shift C_v is used to generate the set of candidate sequences. Given that the number of cell-dependent LP-SS is 4, we suggest to specify 4 different roots.

Proposal 7: Consider 4 different roots for inter-cell interference mitigation.

Regarding the configuration of root and cyclic shifts, the following agreement has been reached

Agreement: **RAN1#119**
CS(s) and/or root(s) used for overlaid OFDM sequence in the time domain are derived from RRC signalling: • FFS: the set of values of CS and root for configuration • FFS: details of the RRC signaling

As mentioned above, the root can simply be specified depending on the physical cell ID N_{ID}^{cell} without need for signaling. Similarly, the CS(s) are derived by the number of configured overlaid OFDM sequences Q and M. Hence, we do not see the need of an explicit signaling of roots and CS(s).

Proposal 8: CS(s) and/or roots(s) are derived from the WUS configuration without need for explicit signaling.

4.2. Single OFDM sequence

It has been agreed to use a ON-sequence configured among a set of specified ON-sequences.

Agreement: **RAN1#118-bis**
In case of overlaid OFDM sequence not carrying information: • Option 1: Single overlaid sequence is on each OOK 'ON' symbol. • Note 1: multiple overlaid OFDM sequences are specified. • Note 2: gNB can configure different overlaid OFDM sequence(s) for different cells.

The configured ON-sequence is the same on all ON-symbols in one cell but may be different in adjacent cells to improve interference randomization. It is our understanding that the single ON-sequence is configured among the set of ON-sequences available if the ON-sequences carry information. For instance, if the maximum number of overlaid OFDM sequences is $Q_{max} = 4$, the configured ON-sequence, if the overlaid OFDM sequences do not carry information, is chosen among those 4 overlaid

OFDM sequences. This avoids specified additional sequences. A natural choice is to use the ZC sequence with $C_v = 0$ (the base overlaid sequence $s(n)$) and a root depending on the neighboring cell IDs.

Proposal 9: Reuse ON-Sequences specified from sequences available if overlaid OFDM sequences carry information.

The LLS in Figure 11 evaluates a *single* overlaid OFDM sequence for three types of receivers (cf. Appendix). As expected, for Manchester Coding $R = 1/2$, there is a ~2dB and ~6dB gain over the ED-WUR for COR-WUR-OOK and COR-WUR, respectively. The larger gain of the COR-WUR is due to the processing gain by carrying out correlations over the entire WUS which requires a more complex receiver.

Applying Pulse Position Coding (PPC) (see Section 3.3) yields a gain of 3dB, 2.8dB and 1.8dB over Manchester Coding for ED-WUR, COR-WUR-OOK and COR-WUR, respectively. The gain for COR-WUR is smaller because the gain does not come from the increased SNR per OOK symbol, but from reduced cross-correlation among the possible transmit signals.

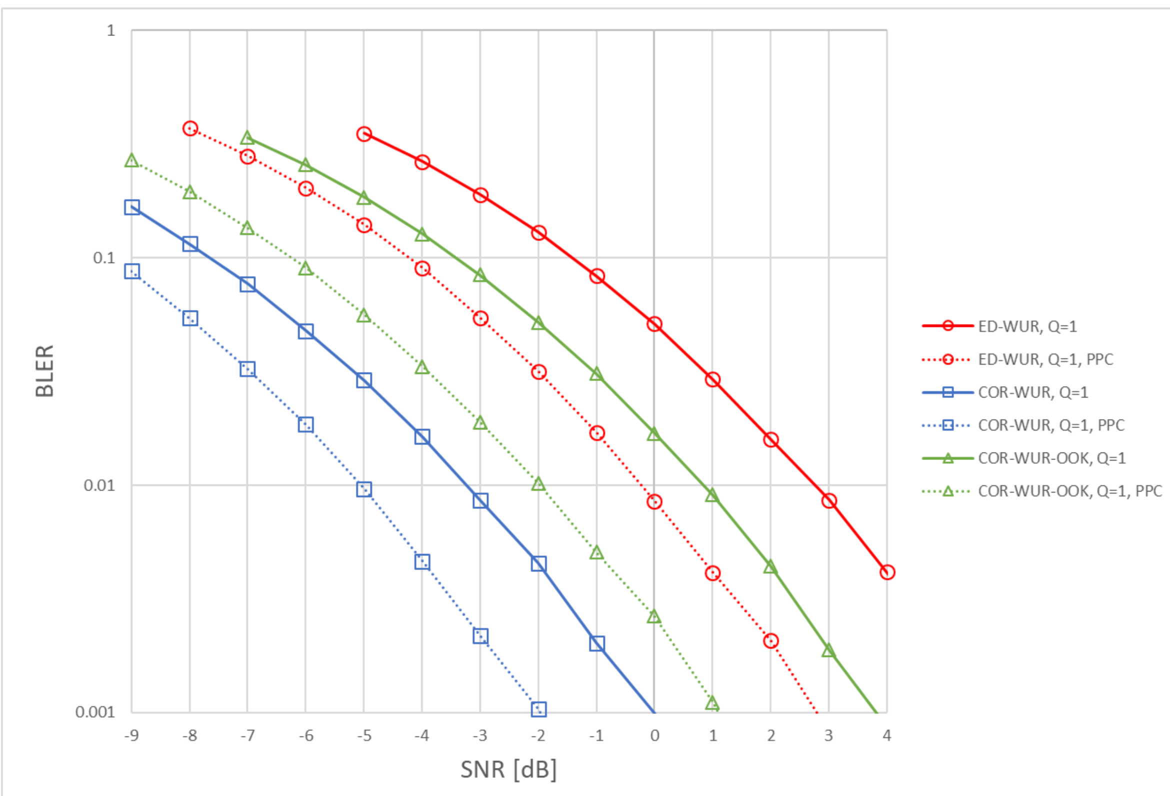

Figure 11: WUR performance in TDL-C with M=4, B=8 and single ON-sequence (Q=1).

Observation 4: Correlation receiver achieves significant gain over energy detection.
Observation 5: For $M = 4$, Pulse Position Coding achieves significant performance gain for all receiver types.

4.3. Multiple OFDM sequences

The relevant agreements from last meetings read:

Agreement: **RAN1#119**

Update the existing agreement as below

In case of overlaid OFDM sequence carrying information, support option 2:

- Option 2: One sequence is selected from multiple candidates overlaid OFDM sequences on each OOK 'ON' symbol, and OFDM-based LP-WUR obtain LP-WUS information at least by overlaid OFDM sequence(s). Consider the following two sub-options for potential down-selection.
  - ~~Option 2-1: The overlaid OFDM sequence(s) carry part of information bits of LP-WUS. OFDM-based LP-WUR can obtain the whole information bits by OFDM sequence(s) and location of the OFDM sequence(s)/OOK 'ON' symbols.~~
  - Option 2-2: The overlaid OFDM sequence(s) carry all information bits of LP-WUS. OFDM-based LP-WUR can obtain the whole information bits by the overlaid OFDM sequence(s)
    - FFS the how the information bits are carried by the overlaid OFDM sequence(s)

Note: the overlaid OFDM sequence in each OOK 'ON' symbol can be different according to information bits to be carried by the overlaid OFDM sequence within the LP-WUS.

Agreement: **RAN1#120**

For WUS information carried by the overlaid OFDM sequence(s), consider at least the following alternatives:

- Alt 1: Raw information bits are mapped to sequence(s)
- Alt 2: Raw information bits are mapped to sequence(s) after channel coding
  - Same channel coding scheme(s) as OOK is applied.
    - FFS same or different rate matching and/or repetition factor as OOK
- Alt 3: Codepoint/Subgroup is mapped to sequence(s)

Denote Q the number of configured overlaid OFDM sequences per OOK ON symbol and $\boldsymbol{s}_q$ the q^{th} sequence, $q = 1,2, \dots Q$. According to our understanding, the raw information bits are $\boldsymbol{b}$, i.e. the input to the channel encoding procedure. In case of codepoint-based mapping of subgroups, $\boldsymbol{b}$ also represents a codepoint.

To better understand the different alternatives, consider the example in Table 6 where 16 subgroups are mapped to 8 codepoints, i.e. a payload size of $B = 3$ bits.

Subgroups	Mapping to payload/codepoint $\boldsymbol{b}$	Output of Channel Coding $\bar{\boldsymbol{d}}$ (no scrambling)
1	{0,0,0}	{0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0,0}
2,3	{0,0,1}	{0,1,0,1,1,0,1,0,0,1,1,1,0,0,0,0,1,0,0,0,1,0,0,1,1,0,1,1,1,1,1,0}
4	{0,1,0}	{1,1,0,0,1,1,0,0,1,0,0,1,0,1,0,1,1,0,1,0,0,1,0,1,1,1,0,1,0,0,1,0}
5,6,7	{0,1,1}	{1,0,0,1,0,1,1,0,1,1,1,0,0,1,0,1,0,0,1,0,1,1,0,0,0,1,1,0,1,1,0,0}
8,9	{1,0,0}	{1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1,1}
10,11,12,13	{1,0,1}	{1,0,1,0,0,1,0,1,1,0,0,0,1,1,1,1,0,1,1,1,0,1,1,0,0,1,0,0,0,0,0,1}
14,15,16	{1,1,0}	{0,0,1,1,0,0,1,1,0,1,1,0,1,0,1,0,0,1,0,1,1,0,1,0,0,0,1,0,1,1,0,1}
1-15	{1,1,1}	{0,1,1,0,1,0,0,1,0,0,0,1,1,0,1,0,1,1,0,1,0,0,1,1,1,0,0,1,0,0,1,1}

Table 6: Example of subgroup to codepoint/payload mapping for 16 subgroups and payload of B=3 bits.

To understand the different mappings, we assume that there is no rate-matching, i.e. $\bar{\boldsymbol{d}} = \boldsymbol{d}$. First consider $Q = 2$ available sequences per OOK ON symbol with possible mappings in Table 7. "OOK" indicates the bits modulated with OOK and transmitted on a single Manchester coded (MC) symbol, containing one OFF and one ON symbol. Here the output of the rate-matched codeword is transmitted. Alt1 maps one bit of the payload $\boldsymbol{b}$ to the sequences. The MC symbol $\bar{d}_1$ already contains information about the codepoint, i.e. it's either codepoints 0,1,6,7 or codepoints 2,3,4,5, which means that b_1 and b_2 are dependent, i.e. if $b_1 = 0$ then $b_2 = 1$ and vice versa. Hence, transmitting b_1 on the first ON symbol and b_2 on the second ON symbol will not give the correlation receiver more information. Therefore, it is best to transmit b_3 in one of the first 2 ON symbols. This enables the correlation receiver to determine the codeword after only 2 ON symbols.

Alt2 proposes to map the channel coded bit $\bar{\boldsymbol{d}}$ to the sequences. As evident from the RM channel code in Table 6, the first B coded bits are sufficient to determine the codepoint. The remaining bits are for redundancy and coding gain which improve FAR. With the same reasoning as in Alt1, it is better to map the bits in cyclically shifted manner to avoid transmitting the same information with OOK and sequence encoding on the same MC symbol.

Alt3 proposes to map one or multiple codepoints (or associated subgroups) to the overlaid sequences. That is, codepoint 1 is signaled on the first ON symbol (remember $Q = 2$), codepoint 2 and the second ON symbol etc. It is clear, that early-termination is not possible since the UE has to wait for the ON symbol that encodes its codepoint. For instance, subgroup 16 has to check ON symbols 7 and 8. It is also unclear how to combine the results of OOK detection with those obtained from the correlations.

ON Symbol Index	1	2	3	4	5	6	7	8	9	…
OOK	$\bar{d}_1$	$\bar{d}_2$	$\bar{d}_3$	$\bar{d}_4$	$\bar{d}_5$	$\bar{d}_6$	$\bar{d}_7$	$\bar{d}_8$	$\bar{d}_9$	…
Alt1	b_3	b_2	b_1	b_3	b_2	b_1	b_3	b_2	b_1	…
Alt2	$\bar{d}_3$	$\bar{d}_1$	$\bar{d}_2$	$\bar{d}_6$	$\bar{d}_4$	$\bar{d}_5$	$\bar{d}_9$	$\bar{d}_7$	$\bar{d}_8$	…
Alt3 (Subgroups)	1	2,3	4	5,6,7	8,9	10,11,12,13	14,15,16	1-16	1	…

Table 7: Example of mapping to multiple OFDM overlay sequences for $Q = 2$ sequences with previous example of mapping for 16 subgroups and payload of $\mathcal{B}$=3 bits.

ON Symbol Index	1	2	3	4	5	6	7	8	9	…
OOK	$\bar{d}_1$	$\bar{d}_2$	$\bar{d}_3$	$\bar{d}_4$	$\bar{d}_5$	$\bar{d}_6$	$\bar{d}_7$	$\bar{d}_8$	$\bar{d}_9$	…
Alt1	b_3b_2	b_1b_3	b_2b_1	b_3b_2	b_1b_3	b_2b_1	b_3b_2	b_1b_3	b_2b_1	…
Alt2	$\bar{d}_3\bar{d}_2$	$\bar{d}_1\bar{d}_4$	$\bar{d}_5\bar{d}_6$	$\bar{d}_7\bar{d}_8$	$\bar{d}_9\bar{d}_{10}$	$\bar{d}_{11}\bar{d}_{12}$	$\bar{d}_{13}\bar{d}_{14}$	$\bar{d}_{15}\bar{d}_{16}$	$\bar{d}_{17}\bar{d}_{18}$	…
Alt3 (Subgroups)	{1}, {2,3}	{4}, {5,6,7}	{8,9}, {10,11,12,13}	{14,15,16}, {1-16}	{1}, {2,3}	{4}, {5,6,7}	{8,9}, {10,11,12,13}	{14,15,16}, {1-16}	{1}, {2,3}	…

Table 8: Example of mapping to multiple OFDM overlay sequences for $Q = 4$ sequences with previous example of mapping for 16 subgroups and payload of $\mathcal{B}$=3 bits.

In the following, we provide out view on how the information bits are carried by the overlaid OFDM sequence(s).

Proposal 10: Support Alternative 1, mapping raw information bits to sequence(s)

4.3.1. Alt 1: Mapping of Raw Information Bits to Sequences

The objective states that the same information $\boldsymbol{b}$ is transmitted irrespectively of the WUR-type, i.e. the information encoded with Q available sequences is part of $\boldsymbol{b}$. Consider the example in Figure 12 for payload size of 8 bits, i.e. $\boldsymbol{b} = [b_0, b_1, \dots, b_7]$, and $Q = 2$ sequences. The figure shows 2 options for encoding $\boldsymbol{b}$ via the 2 sequences, option $\boldsymbol{b}_{S0}$ and $\boldsymbol{b}_{S1}$. The option $\boldsymbol{b}_{S0}$ encodes the same payload as the OOK modulation, i.e. $\boldsymbol{b}_{S0} = \boldsymbol{b}$. Option $\boldsymbol{b}_{S1} = [b_4, b_5, b_6, b_7, b_0, b_1, b_2, b_3]$, i.e. on the first OOK On-symbol bit b_4 is encoded on the next b_5 and so on until the 5th OOK On-symbol where b_0 is encoded. The sequence encoding $\boldsymbol{b}_{S1}$ has the advantage that a correlation-based receiver can decode the entire payload after only 2 OFDM symbols instead of 4 and can go back to sleep earlier. On the other hand, the COR-WUR can also use the entire 4 OFDM symbols for improved decoding performance.

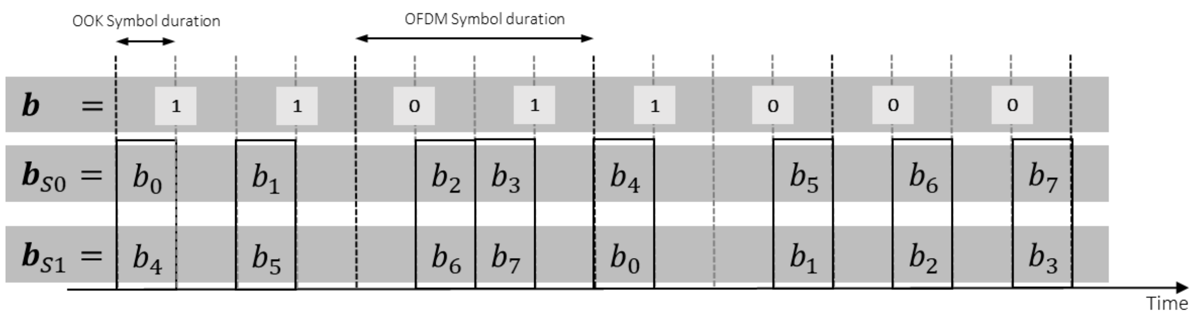

Figure 12: Example of OOK-4 WUS, with M=4, Manchester coding, $Q = 2$ sequences and payload of 8 bits, b = [11011000].

If $Q = 4$ sequences are configured, a possible sequence encoding $\boldsymbol{b}_{S1}$ is shown in Figure 13. Four sequences can encode 2 bits per OOK ON-symbol. This mapping has the following advantages: (1) it avoids encoding the same bits with the sequences as the Manchester coded bit, (2) a correlation-based receiver can decode the entire payload after only 1.5 OFDM symbols and (3) the sequence encoding is repetitive, i.e. the same bits are encoded in the first 2 OFDM symbols as in the last 2 OFDM symbols.

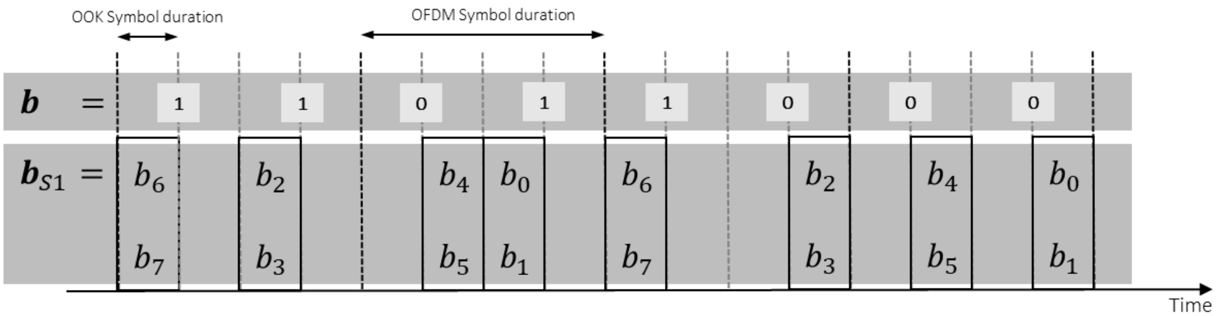

Figure 13:Example of OOK-4 WUS, with M=4, Manchester coding, Q=4 sequences and payload of 8 bits, b = [11011000].

Hence, we propose the following sequence encoding principles:

- The sequence does not encode the same Manchester coded bit
- The encoded is such that a correlation-based receiver can obtain the entire payload in the shortest amount of time
- The sequence encoding is repetitive to facilitate combining and decoding

Figure 14 shows an example with pulse position coding (PPC) and otherwise the same parameters as in the previous example in Figure 12. It can be observed that the position of the ON-sequence within an OFDM symbol encodes 2 bits.

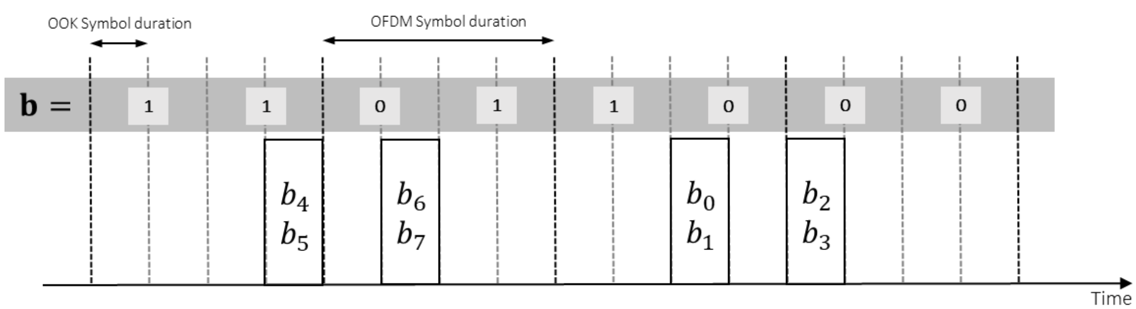

Figure 14: Example of OOK-4 WUS, pulse position coding and M=4, B=8, b = [11011000], L=4 $OFDM\ symbols$ and $Q = 4$ sequences.

Since the power is concentrated in a single OOK-symbol instead of 2 OOK symbols, there is a 3 dB gain for both ED-WUR and COR-WUR-OOK. However, in case of multiple ON-sequences, less bits can be encoded than in the example in Figure 13, because there are fewer ON-pulses. For instance, 2 ON-sequences can encode 4 bits over the entire OOK-4 waveform, as opposed to 8 bits in Figure 13. However, the power in the ON-pulse is larger which results in an overall performance gain.

Figure 15 provides link-level simulation results for multiple overlaid OFDM sequences in various settings. The results labeled “no encoding” use the same bit sequence for the overlaid OFDM sequences as carried by the OOK waveform, e.g. $\boldsymbol{b}_{S0}$ in Figure 12. On the other hand, the results labeled “encoding” use a different bit sequence for the overlaid OFDM sequences, e.g. $\boldsymbol{b}_{S1}$ in Figure 12. It can be observed that COR-WUR achieves a performance gain of ~4.4dB and ~3.3dB over COR-WUR-OOK without encoding and with encoding, respectively, due to the processing gain. For COR-WUR-OOK and COR-WUR encoding yields a 3.3dB and 2dB gain compared to no encoding, respectively.

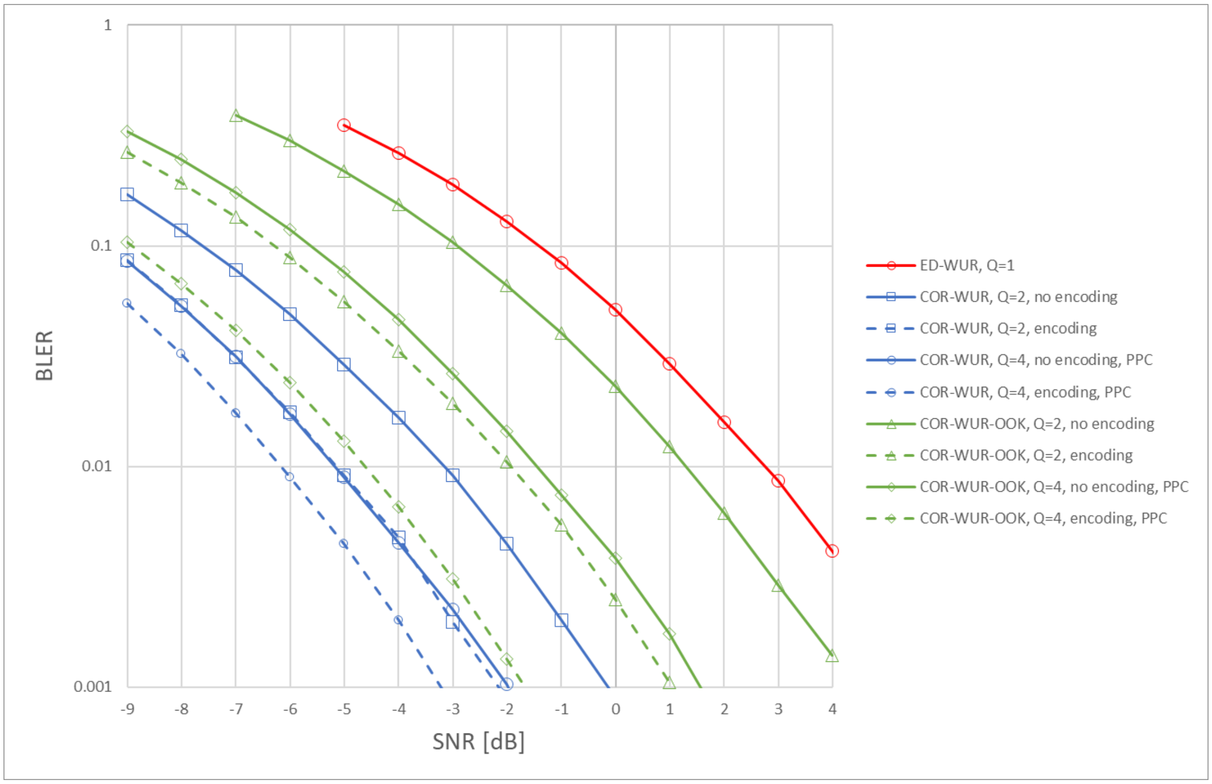

Figure 15: WUR performance in TDL-C with M=4, B=8 and multiple ON-sequences.

With PPC and $Q = 4$ sequences, we observe a 2.8dB and 2.4dB gain for COR-WUR-OOK compared to Manchester coding without and with encoding, respectively. In case of COR-WUR those gains are 2dB and 1dB, respectively.

In summary the following observations are made:

Observation 6:

- **COR-WUR performs better than COR-WUR-OOK due to the processing gain of carrying out longer correlations.**
- **Transmitting the *same* payload as the OOK waveform with the overlaid OFDM sequences but in a *different bit sequence* yields a significant performance gain.**
- **Using Pulse Position Coding and increasing the number of sequences results in a significant performance gain**

Proposal 11: For multiple ON-Sequences, jointly encode the payload with OOK and sequence encoding.

4.3.2. Non-Coherent Time-Overlay Code

In this section, we propose to generate multiple sequences by multiplying base sequences with a modulated symbol. The modulated symbols are obtained by mapping encoded bits to a (complex) constellation point. This method has similarities with the method proposed in Option 4, with the difference that the modulated symbols are the result of a non-coherent encoding procedure.

More precisely, consider L ON-symbols in the WUS transmission which transmits a payload of B bits $\boldsymbol{b} = [b_0, b_1, \dots, b_{B-1}]$. The goal is to encode $\boldsymbol{b}$ using Q OFDM overlay sequences such that the decoding performance is maximized. To this end, we encode $\boldsymbol{b}$ by using a linear non-coherent code $\boldsymbol{G}$ such that

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

where $\boldsymbol{d}$, $\boldsymbol{c}$ and $\boldsymbol{G}$ are the input bits, coded bits $\boldsymbol{c} = [c_0, c_1, \dots, c_{C-1}]$ of length C and the generator matrix, respectively. Subsequently, the modulated N-PSK symbol w_{ml} of message m, $m = 0,1, \dots, 2^B - 1$ and ON-symbol l is obtained by

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

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

$$\boldsymbol{s}_{ml} = \boldsymbol{s}_l \cdot w_{ml}$$

where $\boldsymbol{s}_l$ is the base sequence transmitted on OOK ON-symbol l.

As an example, consider $N = 4$ (e.g. QPSK), $B = 8$ with $R = 1/2$ Manchester Coding, i.e. $L = 8$, a good generator matrix in GF4 is given by

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

The gNB will have pre-stored the codebook of all possible WUS transmissions $\boldsymbol{s}_m$ and the receiver correlates the received signal with all possible transmit messages as detailed in the COR-WUR receiver in Section 6.1.2.2. Note that, the receiver requires joint correlation across the time-overlay duration which also assumes that the channel is coherent during that period.

In order for the code to minimize cross-correlation between all possible messages $\boldsymbol{s}_m$, at least 2 orthogonal base sequences are required.

Figure 16 shows the performance of different encoding schemes for $Q = 2$ base sequences. As in the simulation in Figure 15, the mapping of the bits for sequence encoding has a significant performance impact, about 2dB difference between “encoding” and “no encoding”. By additionally applying the non-coherent code “encoding + time-overlay” we observe an additional 1dB gain. This gain is due to the fact that the code improves cross-correlation between the transmitted codewords. There is no additional computational cost at the transmitter or receiver. Moreover, the time-domain overlay does not impact the performance of the ED-WUR.

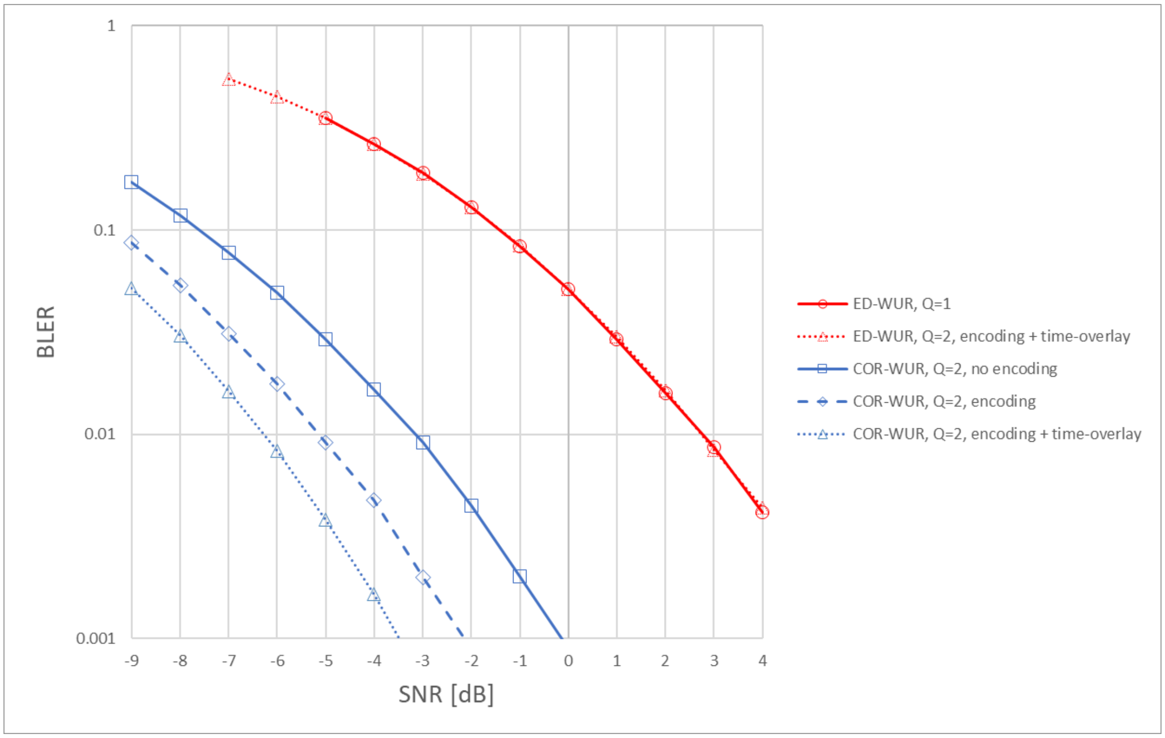

Figure 16: WUR performance in TDL-C with M=4, B=8 and non-coherent encoding scheme.

Observation 7: A time-domain overlay code can significantly improve performance of the overlaid OFDM sequence transmission.

5. LP-SS Design

It has been agreed to support 4 LP-SS to mitigate inter-cell interference.

Agreement:	**RAN1#118**
The number of binary LP-SS sequences is 4.	

It has been agreed to reuse the overlaid OFDM sequences of LP-WUS, i.e. ZC sequences.

Agreement: **RAN1#118-bis**

Support overlaid OFDM sequence(s) for LP-SS:

- LP-SS reuses the overlaid OFDM sequence(s) specified for LP-WUS. The design on overlaid OFDM sequence(s) specified for LP-WUS doesn't target for sync and RRM measurement performance based on overlaid OFDM sequence for LP-SS.
- Whether to transmit LP-SS by using a specified overlaid OFDM sequence is configurable.
    - Applicable at least for OOK-1 and FFS for OOK-4
- From RAN1 perspective, it is not intended to introduce new RAN4 requirements specific to overlaid sequences

The second bullet point is unclear to us. This bullet states that the network can configure the LP-SS such that the gNB transmits a non-specified LP-SS. As far as we understood, the reason behind this is to reuse legacy sync-signals, i.e PSS/SSS or the whole SSB, as overlaid OFDM sequences for LP-SS. The advantage being resource efficiency since LP-SS and SSB can be transmitted on the same resources. However, it is unclear how that could work in practice. Both PSS and SSS are of length 127 with surrounding GBs, the LP-SS is 11 PRBs and hence of length 132 which is more than 127 and this will slightly impact performance. In addition, it is unclear to us how the LP-SS ON-OFF pattern can map to the SSB. The SSB is 4 OFDM symbols long and the middle 11 PRBs are occupied (not to mention the fact that there are no or insufficient GBs in PBCH symbols), i.e. they are ON, which means that the first 4 bits (assuming OOK-1) must map to ON signals. More precisely, the (coded) LP-SS must contain a string of 4 consecutive 1's to be able to contain an SSB. If Manchester coding is used for LP-SS this is impossible.

Another disadvantage concerns the receiver. The receive filter is usually adapted to the spectrum of the known transmitted signal. If now, the transmitted LP-SS is unknown that this receive filter is not well matched to the transmit waveform resulting in a performance degradation.

Moreover, for OOK-4 it is also impossible to use legacy signals since the frequency-domain OOK-4 (M>1) signal does not correspond to any legacy signal.

Observation 8: The receiver should be aware of which LP-SS sequence(s) to expect.

Agreement: **RAN1#118-bis**

To determine the binary sequences for LP-SS, for each M value, down-select the sequence length L from the corresponding candidate values:

- M=1, L= {4,6,8}
- M=2, L= {8,12,16,24}
- M=4, L= {16,24,32,56}
- FFS whether one or more than one L value (s) applied to each of M values M=1,2,4
- FFS the L values applied to other applicable SCS(s)
- FFS the L values if other M values in addition to M=1,2,4 are supported.
- FFS whether the time estimation averaged across multiple LP-SS occasions is applied
- Note 1: with sequence length L, it includes Manchester coding (if supported for LP-SS)
- Note 2: This doesn't preclude any of three agreed sequence types

Above applies at least for both 15kHz and 30kHz.

Agreement:	RAN1#119

For the length L of LP-SS binary sequence, limit the selection to the following:

- M=1, L= {4, 6, 8}
- M=2, L= {8, 12, 16}
- M=4, L= {16, 32}

During RAN1#120, the LP-SS have been further narrowed down to the following (for $M = 4$):

Agreement:	RAN1#120

For M=4, L=16 (if supported), the set of LP-SS sequence is down-selected between:

- Set 1
  [0 1 1 0 1 0 0 1 1 0 1 0 1 0 1 0]
  [0 1 1 0 1 0 1 0 1 0 0 1 1 0 1 0]
  [1 0 1 0 0 1 1 0 1 0 1 0 1 0 0 1]
  [1 0 1 0 1 0 0 1 1 0 1 0 0 1 1 0]
- Set 2
  [1 0 0 0 1 0 0 0 0 0 0 1 0 0 1 0]
  [1 0 0 0 0 1 0 0 1 0 0 0 0 0 1 0]
  [1 0 0 0 0 1 0 0 0 1 0 0 1 0 0 0]
  [1 0 0 0 0 1 0 0 0 0 0 1 0 0 0 1]

Agreement:	RAN1#120

For M=4, L=32(if supported), the set of LP-SS sequence is down-selected between:

- Set 1:
  [0 1 0 1 1 0 1 0 1 0 1 0 1 0 0 1 1 0 1 0 0 1 1 0 0 1 1 0 0 1 0 1]
  [0 1 1 0 0 1 0 1 0 1 1 0 0 1 0 1 1 0 0 1 1 0 1 0 1 0 1 0 0 1 0 1]
  [0 1 0 1 0 1 0 1 1 0 1 0 1 0 0 1 1 0 1 0 1 0 0 1 1 0 1 0 0 1 1 0]
  [0 1 0 1 0 1 1 0 0 1 0 1 1 0 1 0 0 1 1 0 0 1 1 0 1 0 1 0 0 1 0 1]
- Set 2
  [0 0 0 1 1 0 0 0 0 0 0 0 1 0 0 0 1 1 0 0 0 0 0 0 1 0 0 0 1 0 1 0 0]
  [0 0 0 1 0 1 0 0 1 0 0 0 1 0 0 0 0 1 0 0 0 1 0 0 0 0 0 1 0 1 0 0]
  [0 0 1 0 0 0 1 0 1 0 0 0 0 1 0 0 1 0 0 0 0 0 0 1 0 0 1 0 0 0 1 0]
  [0 0 1 0 0 1 0 0 0 0 0 1 1 0 0 0 0 0 1 0 1 0 0 0 0 1 0 0 0 0 1 0]

In the following, we investigate the performance of both options for $L = 16$ and $L = 32$. Set 1 is a Manchester encoded sequence, i.e. for 2 subsequent bits, one bit is zero and the other is one. On the other hand, Set 2 is a sequence encoded with pulse position coding (according to Table 3), e.g. for $L = 16$, bit sequences [1 1 1 1 0 0 0 1], [1 1 1 0 1 1 0 1], [1 1 1 0 1 0 1 1] and [1 1 1 0 0 0 0 0] are encoded with PPC. Similarly, for $L = 32$, bit sequences [0011000011000010], [0010111110100010], [0101111011000101] and [0110001101111001] are encoded with PPC.

As a comparison, we choose to generate the binary LP-SS d_{LP-SS} similarly to the NR PSS, i.e. $d_{LP-SS}(n) = x(m)$ with $m = (n + Gk) mod\ L$, $k = \{0,1,2,3\}$, $G = \lfloor L/4 \rfloor$ and $0 \leq n < (L-1)$. That is, the LP-SS are m-sequences. Subsequently, we use PPC to encode the m-sequences.

For $L = 32$, we obtain the following set of sequences:

- Set 3
  [0 1 0 0 0 0 1 0 0 0 0 1 0 0 0 1 1 0 0 0 0 0 0 0 1 0 0 1 0 1 0 0 0]
  [0 0 0 1 0 0 0 1 1 0 0 0 0 0 0 0 1 0 0 1 0 1 0 0 0 0 0 0 0 1 0 1 0 0]
  [1 0 0 0 0 0 0 1 0 0 1 0 1 0 0 0 0 0 0 0 1 0 1 0 0 0 0 0 0 1 0 0 1 0]
  [0 0 1 0 1 0 0 0 0 0 0 0 1 0 1 0 0 0 0 0 0 1 0 0 1 0 0 1 0 0 0 0 0 0 1]

At the receiver, we assume a correlation detection of the output of the ED over the entire LP-SS and sum the 5 highest peaks. Moreover, for this evaluation, we assume perfect synchronization and CP removal. The transmission power per OFDM symbol is normalized and identical for all schemes.

The result is shown in Figure 17 below.

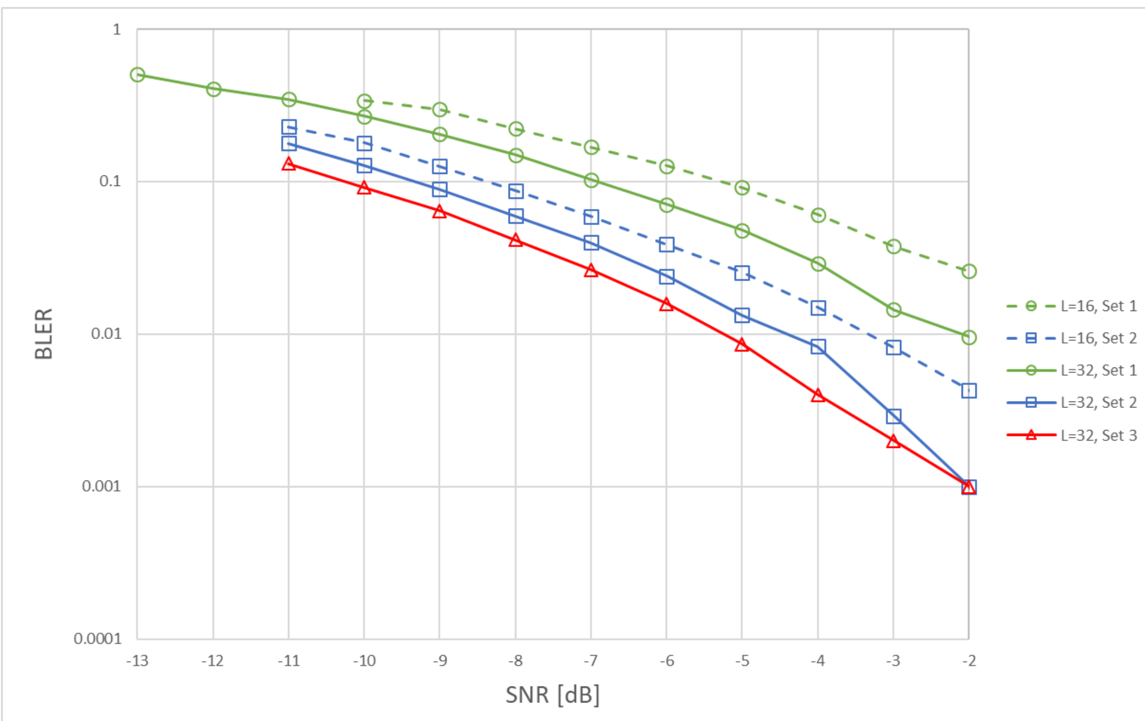

Figure 17: Detection performance for LP-SS transmission, TDL-C, M=4.

It is apparent that Set 2 outperforms Set 1. This can be explained by the "power pooling" effect of PPC, where the power per OFDM symbol is allocated to a single ON symbol instead of 2 ON symbols. In general, it is advantageous to reduce the number of '1's in the binary sequence to concentrate the available power and to improve cross-correlation.

Proposal 12: Support of Pulse Position Coding for LP-SS for $M = 4$.

Interestingly, our proposed Set 3 outperforms Set 2 which indicates that Set 2 can be improved upon.

Observation 9: For $M = 4$ and $L = 32$, the detection performance of Set 2 is not optimal and can be improved.
Proposal 13: For $M = 4$ and $L = 32$, consider Set 3 for superior detection performance.

6. Conclusion

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

Proposal 1: Consider if pulse-shaping is required after sequence design and potential preamble are agreed.

Proposal 2: The DFT-shift is compensated at the LR.

Proposal 3: Do not consider mapping/quantizing WUS in frequency-domain.

Observation 1: Consider a channel encoding procedure for more than 11 information bits, e.g. 16 bits in RRC_CONNECTED mode.

Proposal 4: Specify Manchester Coding as 0 →[1 0] and 1→[0 1].

Observation 2: $M = 4$ with Manchester Coding has the worst coverage compared to $M = 1, 2$.

Proposal 5: For $M = 4$, consider jointly encoding multiple bits into ON pulse position to increase SNR by 3dB.

Observation 3: PAPR increase of Pulse Position Coding for $M = 4$ compared to Manchester encoding depends on the ratio of channel BW to WUS BW and is minor (~0.1dB) for many system configurations.

Proposal 6: Allow configuration of *Pulse Position Coding* for $M = 4$.

Proposal 7: Consider 4 different roots for inter-cell interference mitigation.

Proposal 8: CS(s) and/or roots(s) are derived from the WUS configuration without need for explicit signaling.

Proposal 9: Reuse ON-Sequences specified from sequences available if overlaid OFDM sequences carry information.

Observation 4: Correlation receiver achieves significant gain over energy detection.

Observation 5: For $M = 4$, Pulse Position Coding achieves significant performance gain for all receiver types.

Proposal 10: Support Alternative 1, mapping raw information bits to sequence(s)

Observation 6:

- **COR-WUR performs better than COR-WUR-OOK due to the processing gain of carrying out longer correlations.**
- **Transmitting the *same* payload as the OOK waveform with the overlaid OFDM sequences but in a *different bit sequence* yields a significant performance gain.**
- **Using Pulse Position Coding and increasing the number of sequences results in a significant performance gain**

Proposal 11: For multiple ON-Sequences, jointly encode the payload with OOK and sequence encoding.

Observation 7: A time-domain overlay code can significantly improve performance of the overlaid OFDM sequence transmission.

Observation 8: The receiver should be aware of which LP-SS sequence(s) to expect.

Proposal 12: Support of Pulse Position Coding for LP-SS for $M = 4$.

Observation 9: For $M = 4$ and $L = 32$, the detection performance of Set 2 is not optimal and can be improved.

Proposal 13: For $M = 4$ and $L = 32$, consider Set 3 for superior detection performance.

7. References

[1] RP-213645, "New SID: Study on low-power Wake-up Signal and Receiver for NR", vivo, RAN#94e

[2] TR-38.869, "Study on low-power Wake-up Signal and Receiver for NR", V.18.0.0, 3GPP, 2023

[3] RP-234056, "New WID: Low-power wake-up signal and receiver for NR (LP-WUS/WUR)", CMCC (moderator, RAN VC), RAN#102, Edinburgh, GB, December 11-15, 2023

[4] R1-2405708, "Summary # 5 of discussions on LP-WUS and LP-SS design", Moderator (vivo), RAN1#117

8. Appendix

8.1 Receiver Algorithms

6.1.1. Envelop/Energy Detection in Base-band

The received signal $y(t)$ is first passed through a band-pass filter to extract the WUS frequencies. Subsequently, $y(t)$ is converted into base-band using a sampling frequency $f_{S,WUS}$ resulting in a sampling rate reduction of $r = f_{S,OFDM}/f_{S,WUS}$. Therefore, with K denoting the FFT size of the OFDM transmission, we obtain $N = K/r$ WUS samples per OFDM symbol in time-domain at the receiver base-band. Denote $\boldsymbol{y}_l \in \mathbb{C}^N$ the discrete received signal in base-band of OFDM symbol l. With M the number of OOK symbols per OFDM symbol we have $\boldsymbol{y}_l = [\boldsymbol{y}_{l,1,}\boldsymbol{y}_{l,2} \dots \boldsymbol{y}_{l,M}]$ with $\boldsymbol{y}_{l,i} \in \mathbb{C}^{N_M}$ the $N_M = N/M$ samples per OOK symbol. The energy $e_{i,l}$ of OOK symbol i in OFDM symbol l is then given by

$$e_{i,l} = \sum_{n=1}^{N_M} \left|y_{l,i,n}\right|^2 = \boldsymbol{y}_{l,i}^H \boldsymbol{y}_{l,i}$$

The energy is then compared to a threshold or if Manchester coding is used, the energy values are compared, e.g. for Manchester code $R = 1/2$

$$b_{j,l} = \begin{cases} 0, & if\ e_{i,l} < e_{i+1,l} \\ 1, & if\ e_{i,l} \geq e_{i+1,l} \end{cases}$$

Where $b_{j,l}$ is bit j encoded in 2 consecutive OOK symbols i and $i + 1$ of OFDM symbol l.

6.1.2. Correlation detection in Base-band in Time-Domain

If the ON-sequence $\boldsymbol{a}$ is known the receiver can carry out correlations with all possible transmitted sequences $\boldsymbol{s}_m = [\boldsymbol{s}_{m1}\boldsymbol{s}_{m2} \dots \boldsymbol{s}_{mL}] \in \mathbb{C}^{NL}$. After OFDM demodulation, the WUS signal (sub-carriers) is transformed back into time-domain via Inverse-DFT precoding. We assume that channel estimates are not available, hence a we use a RAKE demodulator for square-law combination of orthogonal signals [Proakis, Figure 14-5-7].

6.1.2.1. Correlation per OOK Symbol

If the correlation is carried out per OOK symbol (referred to as COR-WUR-OOK), the resulting time-domain signal $\boldsymbol{y}_{l,i}$ of the i^{th} OOK symbol is correlated with $\boldsymbol{a}$ and the N_P peaks of the correlator output are combined to exploit frequency-diversity in fading channels i.e.

$$e_{i,l} = \sum_{n=1}^{N_P} \max_{N_P} \left|(\boldsymbol{y}_{l,i} \star \boldsymbol{a})[n]\right|^2$$

where $\max_{N_P}|f(x)|^2$ returns the N_P largest absolute values squared of $f(x)$ and the linear cross-correlation function $(f \star g)[n]$is defined as

$$(f \star g)[n] = \sum_{m=-N}^{N} f^*[m]\, g[m+n]$$

6.1.2.2. Correlation per WUS

If the correlation is carried out over the entire WUS to achieve the largest processing gain (referred to as COR-WUR), the message $m, m = 0,1, \dots, 2^B - 1$, for payload of B bits WUS is decoded as

$$\hat{m} = \underset{m}{\operatorname{argmax}} \sum_{n=1}^{N_P} \max_{N_P} |(\boldsymbol{y} \star \boldsymbol{s}_m)[n]|^2$$

where $\boldsymbol{y}$ is the time-domain WUS and $\boldsymbol{s}_m$ is the known WUS for message m.

6.2. Simulation Assumptions

Unless otherwise stated, the link-level simulation assumption in Table 9 are used.

Parameter	Value
Carrier Frequency	2.6 GHz (FDD)
Waveform	OOK-4
Channel Structure	Option 3: Payload only (no CRC)
SCS	30 kHz
WUS payload	8 bits
Configuration of LP-WUS Signal	OOK-4: M=4 ON-Sequence = Zadoff-Chu Multiple Sequences are obtained via different cyclic shifts
WUS Duration	4 OFDM symbols

Code Scheme	Manchester Code R=1/2 with transmission of encoded bits
Channel BW	20MHz (51 PRBs @ 30kHz SCS)
LP-WUS BW	5 MHz (14 PRBs = 168 SCs) 148 SCs for WUS (4.44 MHz) + 10 SCs GB on each side
Filter	3rd order Butterworth with 4.32 MHz BW
Adjacent Sub-carrier Interference (ACI)	Random 64-QAM symbols with 0dB
WUS Sampling Rate	7.68 MHz
ADC bit-width	inf
Channel Model	TDL-C, 300ns Delay Spread
Timing Error	0
Frequency Error	0
Antenna configuration	1Tx, 1Rx
UE speed	0 km/h
Receiver	Energy Detector, Correlation Detector (Energy of 5 highest peaks is combined)

Table 9: Link-level simulation assumptions.