
Agenda Item: 9.6.1
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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 Y_{ms} 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 \mathbf{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 $\mathbf{b} = [b_0, b_1, \dots, b_{B-1}]$ are encoded and the resulting C coded bits $\mathbf{c} = [c_0, c_1, \dots, c_{C-1}]$ 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 \mathbf{S}_m of message $m = 0, 1, \dots, 2^B - 1$ is mapped to the overall resources \mathbf{X}_m of N sub-carriers and OFDM-modulated resulting in the time-domain signal $\mathbf{x}_m(t)$.

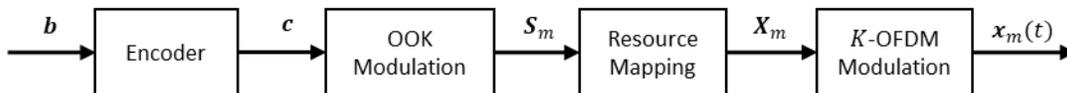


Figure 1: Overall transmission block-diagram per OFDM symbol.

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 \mathbf{s}_m in time-domain for message $m = 0, 1, \dots, 2^C - 1$ as

$$\mathbf{s}_m = [s_0, s_1, \dots, s_{C-1}]$$

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

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

Hereby, the ON-signal \mathbf{a} can be *any complex sequence* and can be different for every s_i .

In case of multiple ON-sequence, consider Q ON-sequences \mathbf{a}_q able to encode $B_2 = \log_2 Q$ bits.

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

The simplest OOK scheme allocates the ON-signal \mathbf{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.

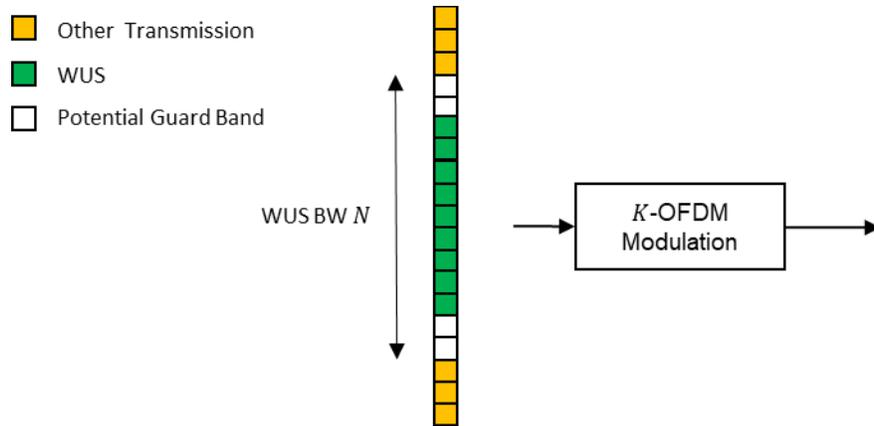


Figure 2: Block-diagram for OOK-1.

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 s_i of length N'_B , the sequence is 0 if the corresponding bit is zero and $s_i = \mathbf{a}$ otherwise. The resulting sequence is DFT precoded and the output is mapped to the overall transmission bandwidth.

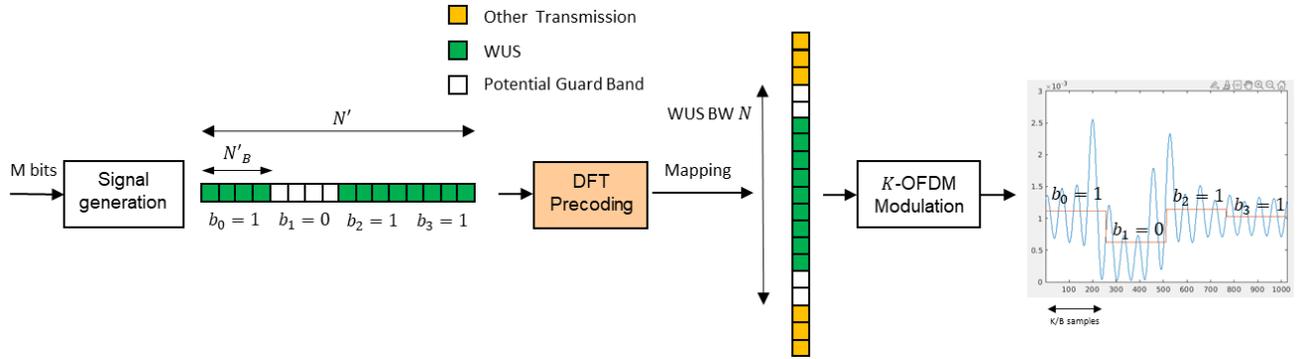


Figure 3: OOK-4, Block-diagram with $M=4$.

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.

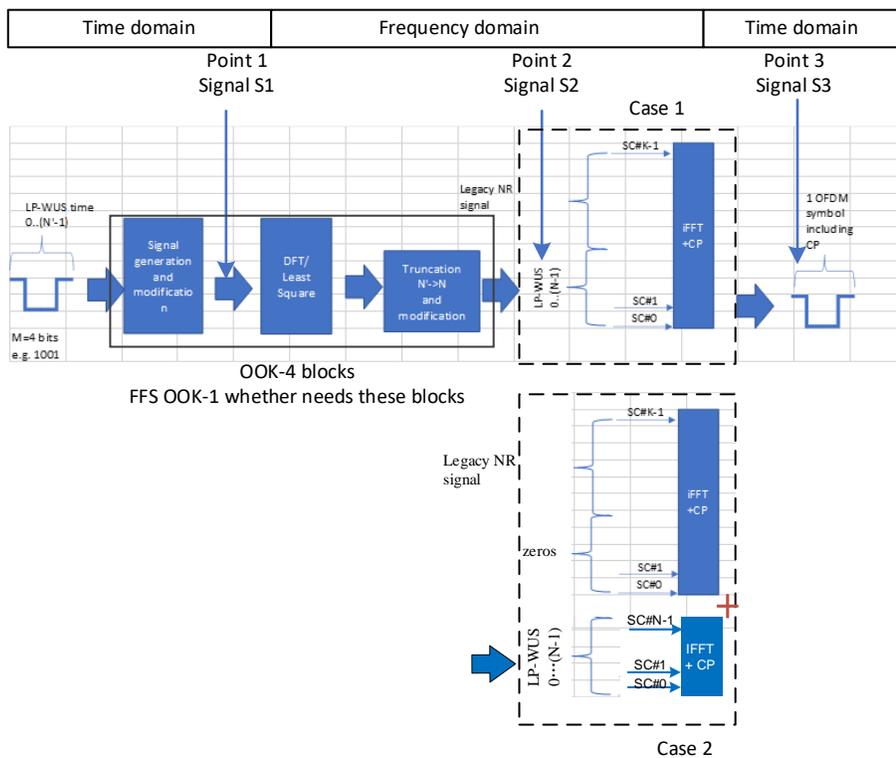


Figure 4: Block-diagram for LP-WUS waveform generation.

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.

2.4.4. LP-WUS Multiplexing Before or After IFFT

It has been proposed to multiplex LP-WUS with NR in time-domain to reduce interference with legacy NR transmission. We do not see any advantage in time-multiplexing the two transmissions. The resource overhead due to the additional CP will be approximately the same as the required guard-bands when the two transmissions are multiplexed in frequency-domain. Simulations have shown that with sufficient GBs, the impact of NR interference is negligible. On the other hand, TDM requires two processing chains at the gNB which increases complexity.

Proposal 4: Multiplexing NR and WUS in frequency-domain is the base line.

3. Overlaid OFDM Sequences

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

3.1. 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

During the meeting RAN1#118-bis, it has been agreed as a working assumption, that OOK-1 is a special case of OOK-4 with $M = 1$, i.e. the ON-sequences are specified in time-domain. This working assumption allows for a unified design and specification on the overlaid sequences in time-domain but has not been confirmed yet due to concerns about the DFT size of 2^n . We support this working assumption but propose to discuss the DFT size separately.

Working Assumption

RAN1#118-bis

For overlaid OFDM sequences for LP-WUS in time or frequency domain, down-select between following two options for OOK-1 and OOK-4 $M=1$

- Option 1-1(compromised): overlaid sequence(s) are the sequence(s) of an OOK on symbol before DFT/LS processing for OOK-4 with $M=1$.
 - ZC sequence in time domain \rightarrow DFT/LS \rightarrow IFFT to result in a frequency domain sequence that is ZC sequence or a sequence that is close to ZC sequence, no additional operation
 - It doesn't preclude additional operation for OOK-4 with $M > 1$.
 - DFT size is 2^n

Note1: OOK-1 is considered as specific case of OOK-4 with $M=1$.

Note: Concerns were raised by Ericsson and CATT that the above is not beneficial in terms of implementation complexity.

Proposal 5: Confirm working assumption and agree Option 1-1 for OOK-4 ($M=1$)/OOK-1.

From previous agreements, the BW of LP-WUS is $X = 11$ PRBs which means 132 REs. The question is now how to define the ZC sequences in time-domain, the length of the DFT and the mapping to the LP-WUS REs.

$$X_q(m) = e^{-j\frac{\pi qm(m+1)}{B_{ZC}}}, m = 0, \dots, B_{ZC} - 1$$

- M=1, 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 $s(n)$ is generated by extension of X_q

$$s(n) = X_q(n \bmod B_{ZC}), n = 0, \dots, L_{ZC} - 1$$

- With CS(s) applied to the base overlaid OFDM sequence if any: C_v denotes the potential cyclic shift (s)

$$s'(n) = X_q((n + C_v) \bmod B_{ZC}), n = 0, \dots, L_{ZC} - 1,$$

- M=2, 4, B_{ZC} is down-selected from the following:

- Alt1: 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 $s(n)$ is generated by extension of X_q

$$s(n) = X_q(n \bmod B_{ZC}), n = 0, \dots, L_{ZC} - 1$$

- With CS(s) applied to the base overlaid OFDM sequence if any: C_v denotes the potential cyclic shift (s)

$$s'(n) = X_q((n + C_v) \bmod B_{ZC}), n = 0, \dots, L_{ZC} - 1$$

- Note it doesn't preclude any pulse shaping scheme if any.

- Alt2: B_{ZC} is given by the largest prime number such that $B_{ZC} \leq L_{ZC} - G$, L_{ZC} is the overlaid OFDM sequence length.

- The base overlaid sequence $s(n)$ is generated by inserting zeros before and/or after X_q . The total number of zeros is $L_{ZC} - B_{ZC}$.

$$\text{For example, } s(n) = [0, \dots, 0, X_q(0), \dots, X_q(B_{ZC} - 1), 0, \dots, 0],$$

$$n = 0, \dots, L_{ZC} - 1$$

- With CS(s) applied to the base overlaid OFDM sequence if any: C_v denotes the potential cyclic shift (s)

$$\text{For example, } s'(n) = [0, \dots, 0, Y_q(0), \dots, Y_q(B_{ZC} - 1), 0, \dots, 0],$$

$$n = 0, \dots, L_{ZC} - 1$$

$$\text{where } Y_q(j) = X_q((j + C_v) \bmod B_{ZC}),$$

$$j = 0, \dots, B_{ZC} - 1$$

- FFS the value of G, with $G > 0$.

- FFS value(s) of root q and/or CS C_v for generating the candidate sequences if applied

- FFS on whether changes are needed to handle intra-/inter-cell interference

- **Note: the overlaid OFDM sequence in time domain is based on potential modification to ZC sequence as listed above.**

- **Above overrides any previous RAN1 agreement / working assumption**

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 , which is generate as

$$s'(n) = X_q((n + C_v) \bmod B_{ZC})$$

with

$$X_q(m) = e^{-j \frac{\pi q m(m+1)}{B_{ZC}}}$$

where $m = 0, 1, \dots, B_{ZC} - 1$, $n = 0, 1, \dots, L_{ZC} - 1$, $B_{ZC} = PP(L_{ZC})$ with $PP(x)$ denoting the largest prime less or equal to x , q is the root and C_v is the cyclic shift. If there is only a single sequence, we have $C_v = 0$ and $s(n) = s'(n)$.

For $M = 2, 4$, there is a second alternative that does not use a cyclic extension of the ZC sequence. Instead, it adds zeros to the beginning and end of the prime length ZC sequence to obtain a sequence of length L_{ZC} . These zeros will then be interpolated by the DFT processing into a frequency-domain sequence. However, the inserted zeros lead to shorter sequence compared to Alt1 which decreases the correlation performance. Moreover, it is desirable to have identical signal generation across all M . Hence, we prefer Alternative 1.

Proposal 6: Select Alternative 1 for Zadoff-Chu time-domain signal generation.

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

Agreement:	RAN1#119
<p>At least for $M > 1$, for the overlaid OFDM sequence in time domain, down-select the sequence length L_{ZC} from the following:</p> <ul style="list-style-type: none"> • Alt1: $L_{ZC} = 2^k / M$, with $2^k \leq 12 * X$ <ul style="list-style-type: none"> ○ FFS how to map the generated overlaid OFDM sequence to the frequency domain • 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) 	

Alternative 1 proposes to choose a sequence length that is a power of 2, whereas the sequence length in Alt2 is equal to number of sub-carriers available. The main reason for Alt1 is that the DFT processing is much more efficient when the input length is a power of 2 which is desirable if the sequences are not pre-computed and stored in memory. As an example, assume $M = 4$ and $Q_{max} = 4$ overlaid sequences, which means that 64 different (frequency-domain) sequences of length 132 have to be stored. Assuming a 16-bit resolution per real number, one sequence requires 4224 bits of storage and hence 64 sequences require 33 Kbytes. Smaller values of M presumably require less sequences/memory. We think that it is reasonable to pre-compute the sequences. Moreover, longer sequences perform better and thus, we favor Alt2.

Proposal 7: For the length of the overlaid OFDM sequence in time-domain, select Alternative 2.

With Alternative 2, 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#118-bis
Regarding the maximum number of candidate overlaid sequences to carry LP-WUS information per OOK ON chip for one cell, down-select from the below:	
<ul style="list-style-type: none"> • 4 • 8 • 16 • 32 • 64 for M not larger than 2. • FFS whether the same or different set of candidate overlaid sequences are used for different cells. 	

Agreement:	RAN1#119
Regarding the maximum number of candidates overlaid sequences to carry LP-WUS information per OOK ON chip for one cell:	
<ul style="list-style-type: none"> • support maximum 4 candidates overlaid sequences for M=4 	

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 $i = 0, 1, \dots, Q_{max} - 1$. 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}	$Q = 2$	$Q_{max} = 4$	$Q_{max} = 8$	$Q_{max} = 16$	$Q_{max} = 32$	$Q_{max} = 64$
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}	{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,

					100, 104, 108, 112, 116, 120, 124}	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 1: 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 1, the potential amount of sequences that need to be stored at the gNB and UE is large. However, the number can be reduced, by 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.

In summary, it is sensible to define the maximum number of candidate sequences as a function of M . We propose $Q_{max} = \{16, 8\}$ for $M = \{1, 2\}$, respectively.

Proposal 8: Consider the number of maximum candidate sequences to be a function of M . Longer sequences allow for a larger number of candidate sequences with the same cyclic shifts.

Proposal 9: Consider 16 and 8 as maximum number of candidate sequences for M of 1 and 2, respectively.

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 10: 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 11: CS(s) and/or roots(s) are derived from the WUS configuration without need for explicit signaling.

3.2. 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 12: Reuse ON-Sequences specified from sequences available if overlaid OFDM sequences carry information.

The LLS in Figure 5 evaluates a *single* overlaid OFDM sequence for three types of receivers (cf. Appendix). As expected, for Manchester Coding $R = 1/2$, there is a ~ 2 dB and ~ 6 dB 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 4.2) 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.

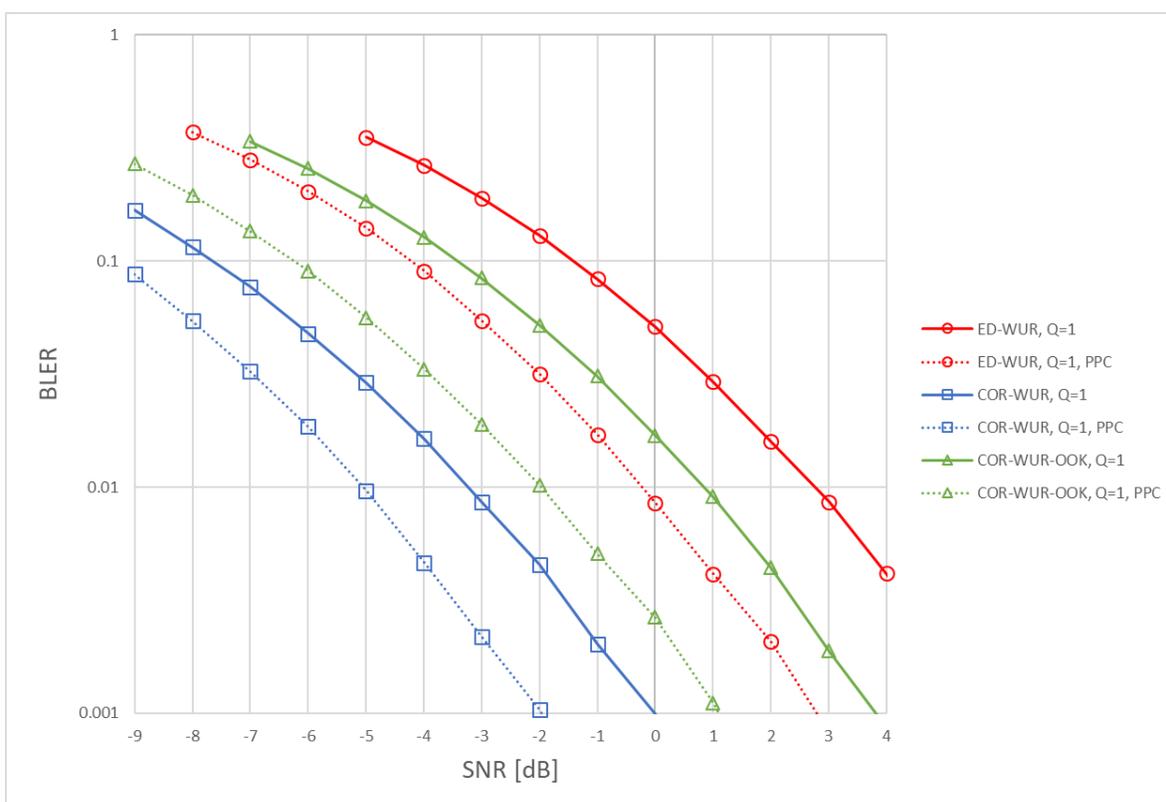


Figure 5: WUR performance in TDL-C with $M=4$, $B=8$ and single ON-sequence ($Q=1$).

Observation 1: Correlation receiver achieves significant gain over energy detection.

Observation 2: For $M = 4$, Pulse Position Coding achieves significant performance gain for all receiver types.

3.3. Multiple OFDM sequences

The updated agreement from last meeting reads:

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.

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

3.3.1. Option 2: Mapping of WUS Information to Sequences

The objective states that the same information \mathbf{b} is transmitted irrespectively of the WUR-type, i.e. the information encoded with Q available sequences is part of \mathbf{b} . Consider the example in Figure 6 for payload size of 8 bits, i.e. $\mathbf{b} = [b_0, b_1, \dots, b_7]$, and $Q = 2$ sequences. The figure shows 2 options for encoding \mathbf{b} via the 2 sequences, option \mathbf{b}_{S0} and \mathbf{b}_{S1} . The option \mathbf{b}_{S0} encodes the same payload as the OOK modulation, i.e. $\mathbf{b}_{S0} = \mathbf{b}$. Option $\mathbf{b}_{S1} = [b_4, b_5, b_6, b_7, b_0, b_1, b_2, b_3]$, 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 \mathbf{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.

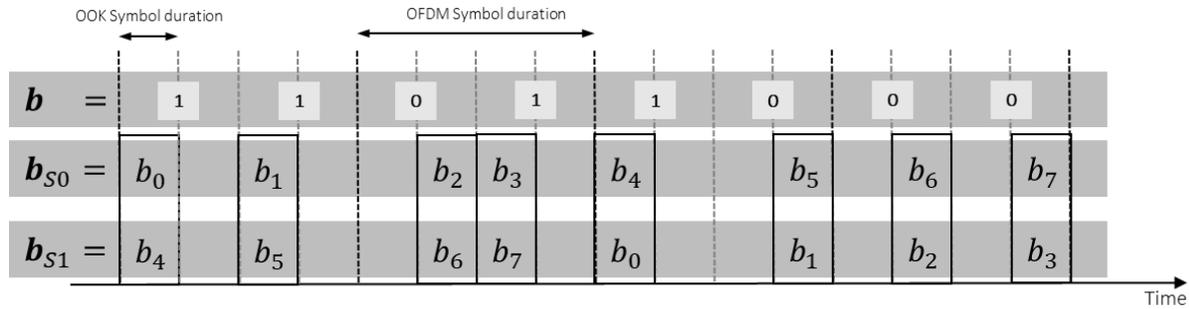


Figure 6: Example of OOK-4 WUS, with $M=4$, Manchester coding, $Q = 2$ sequences and payload of 8 bits, $b = [11011000]$.

If $Q = 4$ sequences are configured, a possible sequence encoding b_{S1} is shown in Figure 7. 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.

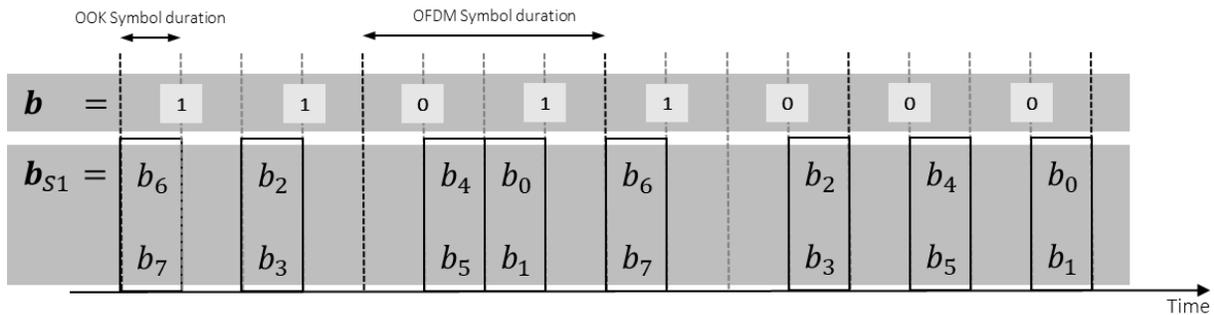


Figure 7: Example of OOK-4 WUS, with $M=4$, Manchester coding, $Q=4$ sequences and payload of 8 bits, $b = [11011000]$.

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 8 shows an example with pulse position coding (PPC) and otherwise the same parameters as in the previous example in Figure 6. It can be observed that the position of the ON-sequence within an OFDM symbol encodes 2 bits.

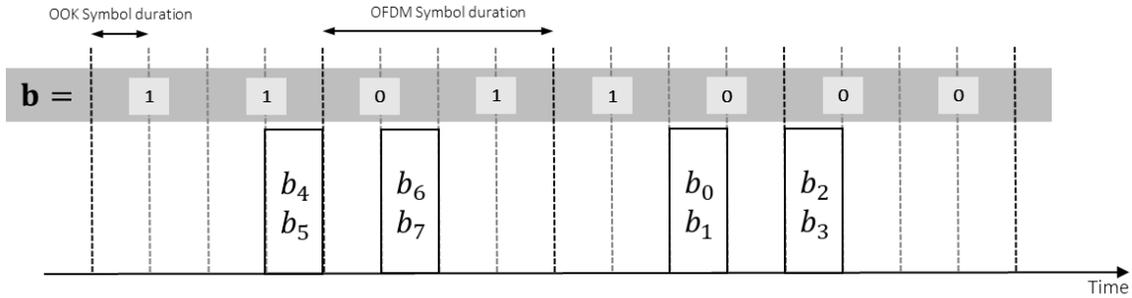


Figure 8: Example of OOK-4 WUS, pulse position coding and $M=4$, $B=8$, $b = [11011000]$, $L=4$ OFDM symbols and $Q = 4$ sequences.

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 7, 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 7. However, the power in the ON-pulse is larger which results in an overall performance gain.

Figure 9 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. b_{S0} in Figure 6. On the other hand, the results labeled “encoding” use a different bit sequence for the overlaid OFDM sequences, e.g. b_{S1} in Figure 6. It can be observed that COR-WUR achieves a performance gain of ~ 4.4 dB and ~ 3.3 dB 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.

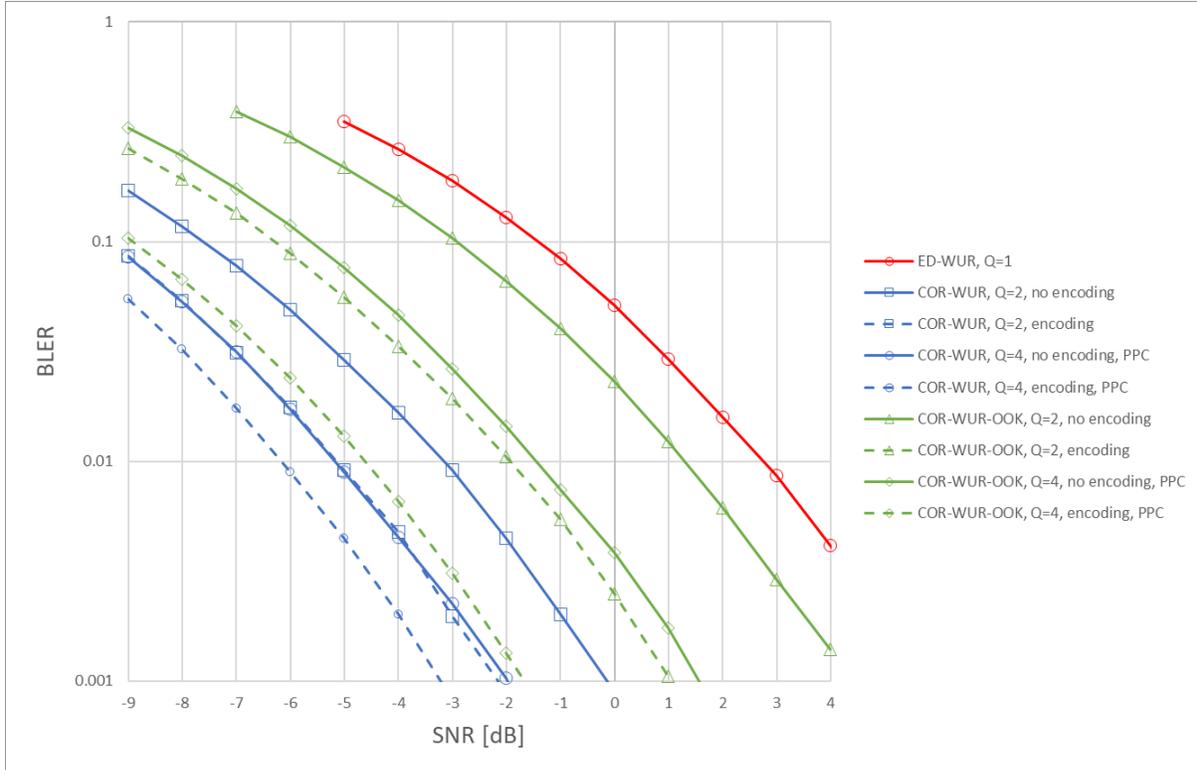


Figure 9: WUR performance in TDL-C with $M=4$, $B=8$ and multiple ON-sequences.

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 3:

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

3.3.2. 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 $\mathbf{b} = [b_0, b_1, \dots, b_{B-1}]$. The goal is to encode \mathbf{b} using Q OFDM overlay sequences such that the decoding performance is maximized. To this end, we encode \mathbf{b} by using a linear non-coherent code \mathbf{G} such that

$$\mathbf{c} = \mathbf{d}\mathbf{G}$$

where \mathbf{d} , \mathbf{c} and \mathbf{G} are the input bits, coded bits $\mathbf{c} = [c_0, c_1, \dots, c_{C-1}]$ 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 \mathbf{c} . Finally, the transmitted sequence \mathbf{s}_{ml} is given by

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

where \mathbf{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

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

The gNB will have pre-stored the codebook of all possible WUS transmissions \mathbf{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 \mathbf{s}_m , at least 2 orthogonal base sequences are required.

Figure 10 shows the performance of different encoding schemes for $Q = 2$ base sequences. As in the simulation in Figure 9, 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.

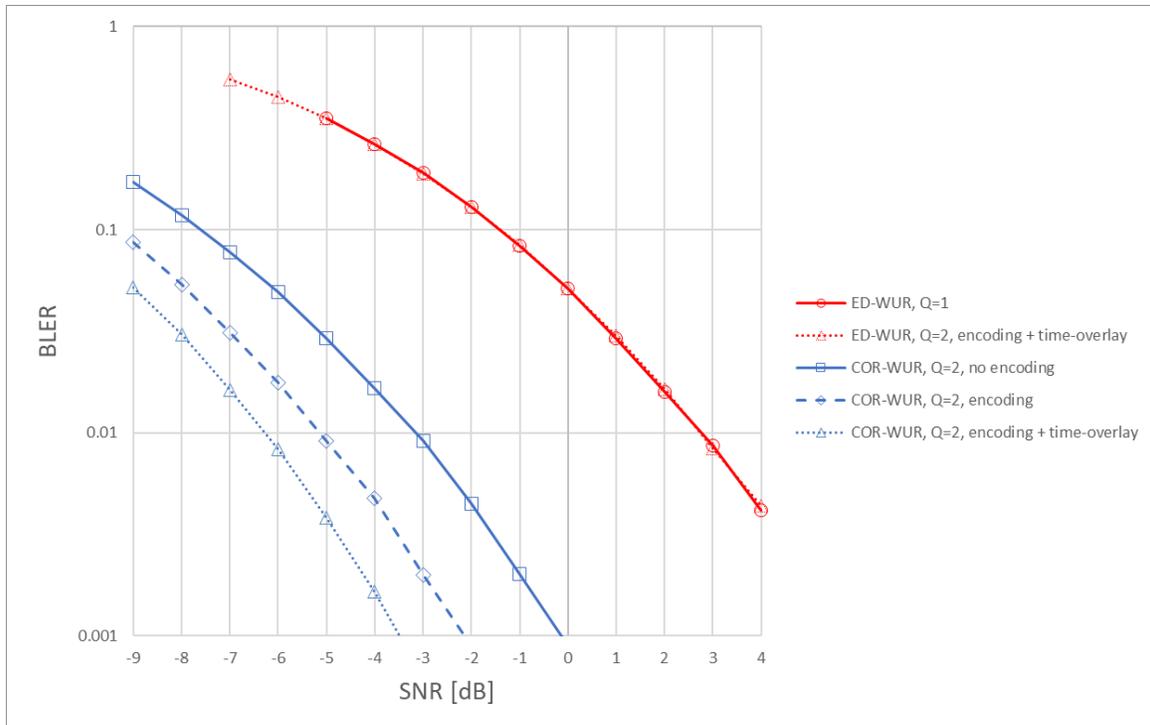


Figure 10: WUR performance in TDL-C with $M=4$, $B=8$ and non-coherent encoding scheme.

Observation 4: A time-domain overlay code can significantly improve performance of the overlaid OFDM sequence transmission.

4. Coding

The study on LP-WUS [2] suggests to utilize Manchester Coding, where one input bit is mapped to two coded bits, e.g. 0->{1,0} and 1->{0,1}. 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.

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

Agreement:

RAN1#118-bis

Down-select among the following alternatives for LP-WUS information for OOK to meet target requirements:

- adding CRC
- channel coding other than Manchester coding, including rate matching if any
- mapped to a binary sequence/codeword, include truncation if any

All of the above alternatives aim to add redundancy to the information bits to increase error protection. Obviously, this comes at the cost of increased resource overhead, e.g. longer sequences or additional bits for CRC.

Subsequently, we propose a line or channel 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.

4.1. Discussion on Manchester Coding

During the last meeting, the following agreement has been made

Agreement:	RAN1#118
Support Manchester coding for LP-WUS	
<ul style="list-style-type: none">• FFS other coding schemes.	

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

1: $0 \rightarrow [0\ 1]$ and $1 \rightarrow [1\ 0]$ or

2: $0 \rightarrow [1\ 0]$ and $1 \rightarrow [0\ 1]$

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 \rightarrow chips {10}, bit 1 \rightarrow chips {01}
- FFS: Variant of the above for CP handling

Therefore, we suggest to follow this MC mapping.

Proposal 14: Specify Manchester Coding as $0 \rightarrow [1\ 0]$ and $1 \rightarrow [0\ 1]$.

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.

4.2. 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 $\mathbf{b} = b_0, b_1, \dots, b_{B-1}$ are encoded to C output bits $\mathbf{c} = c_0, c_1, \dots, c_{C-1}$ according to

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

where $m = 0, 1, \dots, 2^B - 1$ is the message and \bar{c} is the *decimal* representation of codeword 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 $R = 1/2$.

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 $R = 1/2$.

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 11 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.

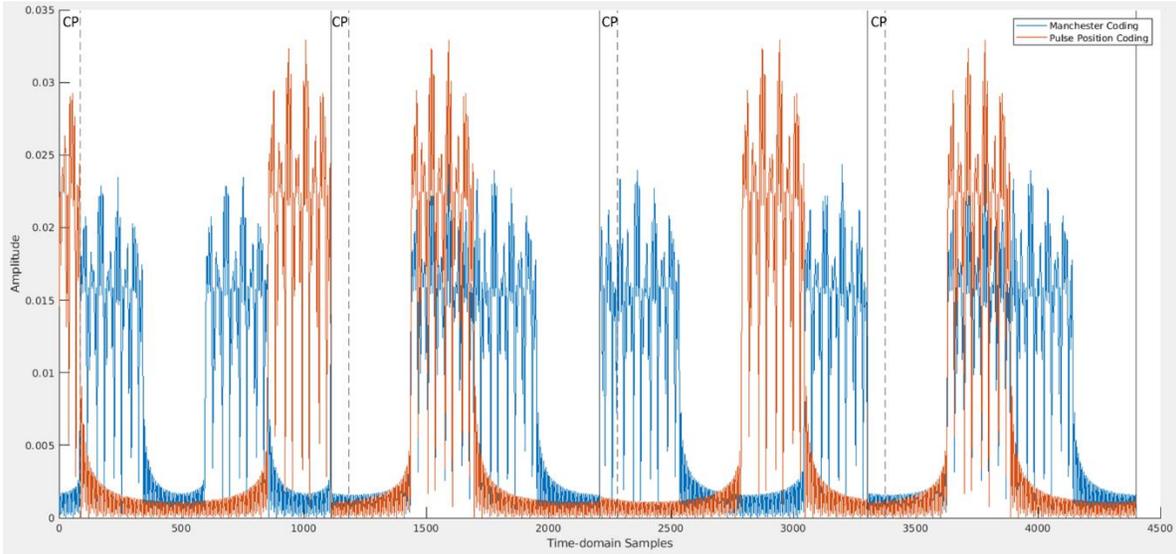


Figure 11: Transmit Signal for payload $\mathbf{b} = [1\ 1\ 0\ 1\ 1\ 0\ 0\ 1]$, $M = 4$, 4 OFDM symbols, no ACI.

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

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

4.2.1. 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 that 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.

4.2.2. 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 5 are used. The transmission power of the WUS per OFDM symbol is kept constant. For an ON-sequence $\mathbf{a} = [a_0 a_1 \dots a_{N'_B-1}]$ of length N'_B , a *truncated Zadoff-Chu* sequence is used which is generate as

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

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 11, 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 12. The PAPR increases for MC and PPC to 9.3dB and 11.2dB, respectively.

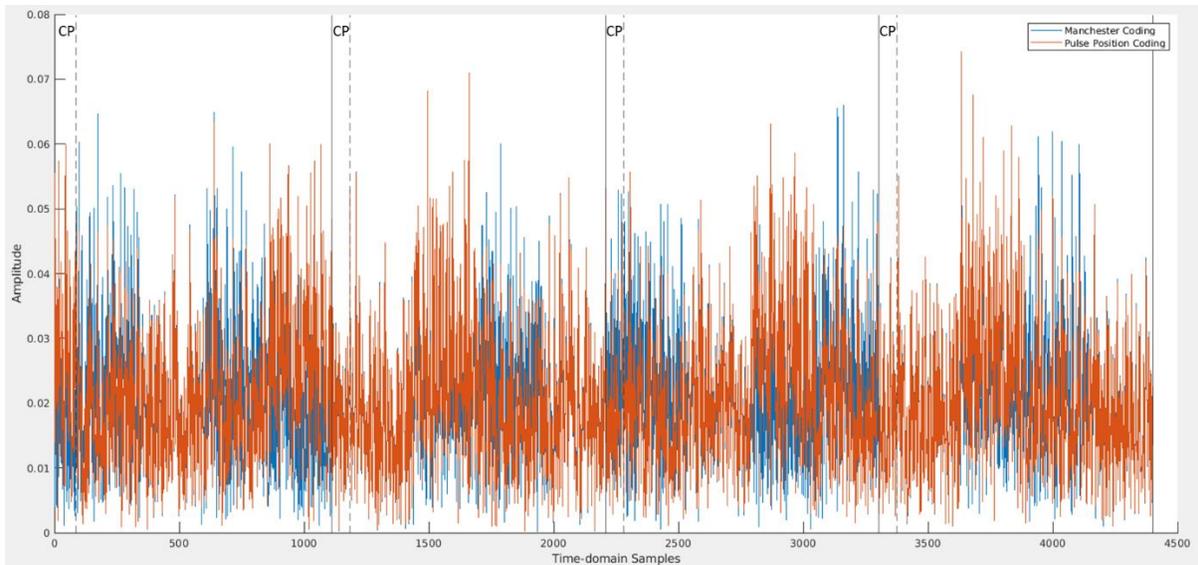


Figure 12: Transmit Signal for payload $\mathbf{b} = [11011001]$, $M=4$, 4 OFDM symbols, with 64-QAM ACI.

Figure 13 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.

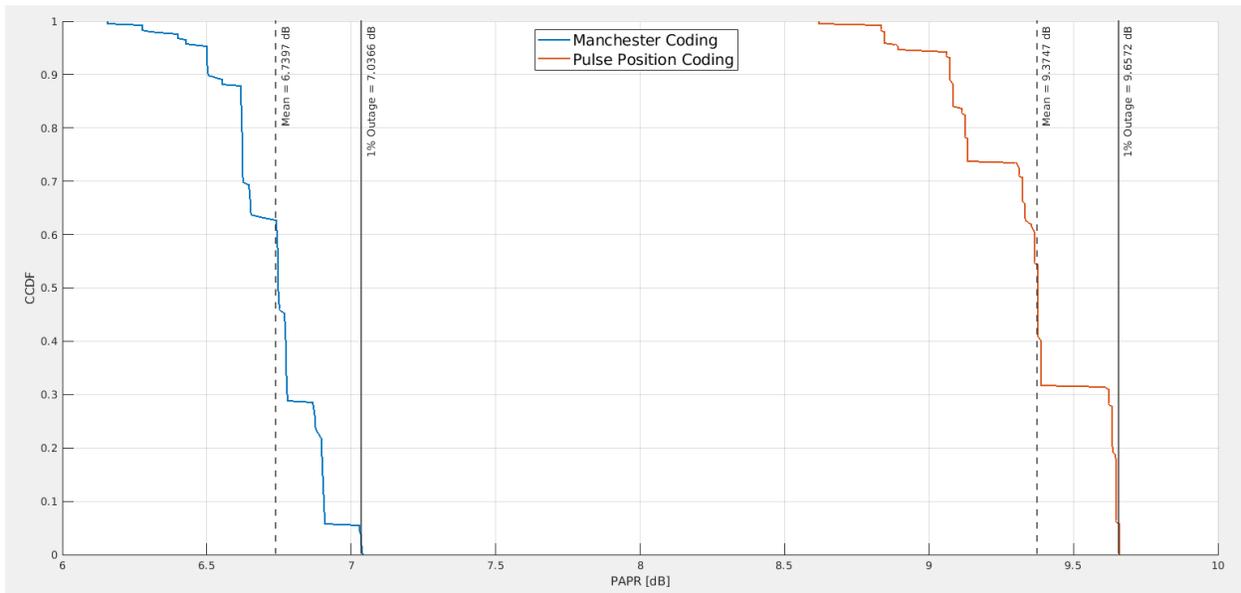


Figure 13: CCDF of PAPR, no ACI, 20MHz, 30kHz SCS, M=4, 8 bit payload, 4 OFDM symbols, average over 10k realizations.

Figure 14 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.

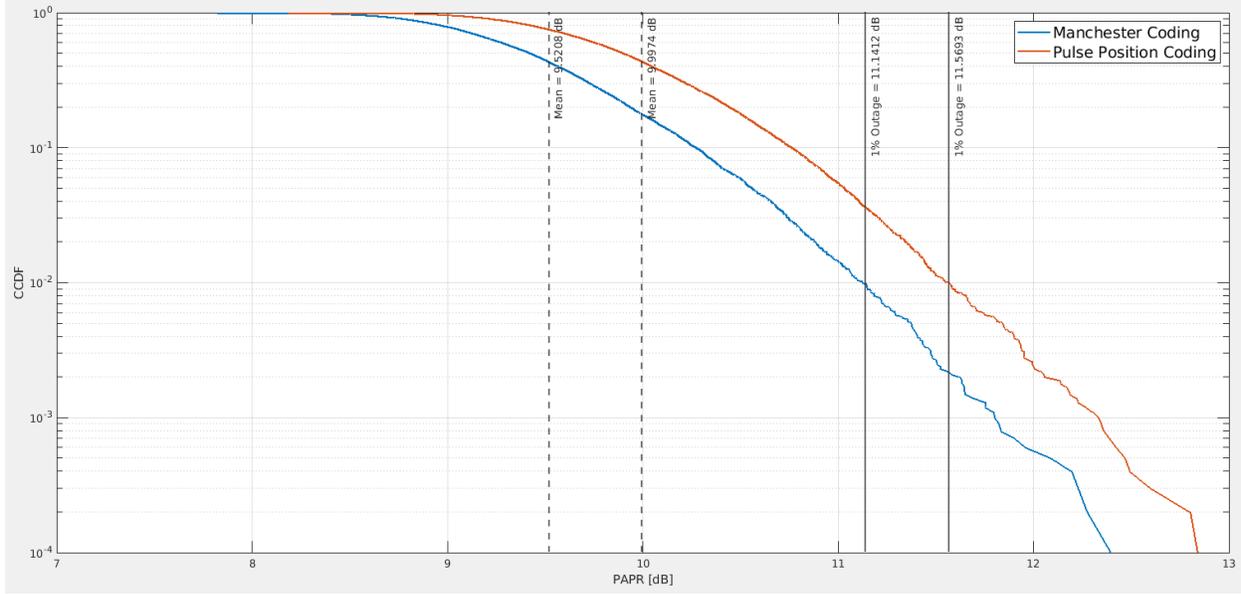


Figure 14: CCDF of PAPR, 64-QAM ACI, 1PRB GB, 20MHz, 30kHz SCS, M=4, 8 bit payload, 4 OFDM symbols, average over 10k realizations.

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 6: 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 (~ 0.1 dB) for many system configurations.

Figure 15 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.

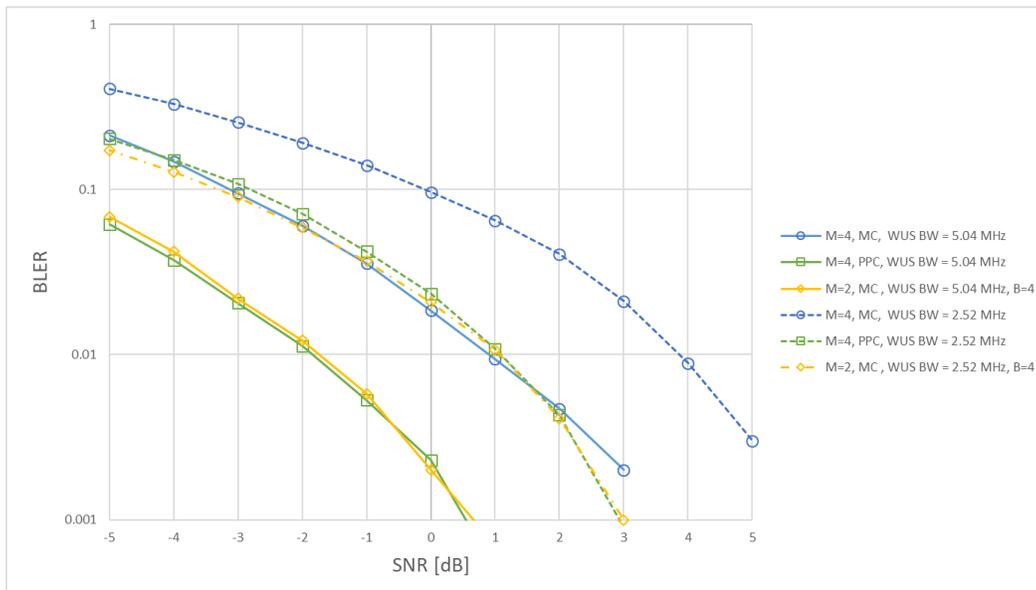


Figure 15: BLER vs. SNR, ED-WUR, 20 MHz (106 PRBs), SCS=15 kHz, M=4, 8 bit payload.

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

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 7: 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}

In the following, we investigate if coding impacts the detection performance of the LP-SS. The general simulation assumptions are detailed in Table 5. In particular, we choose $M = 4$ and a LP-SS duration of one slot, i.e. 14 OFDM symbols. Therefore, the LP-SS is of length $L = 56$ and $L = 28$ OOK symbols for uncoded and coded (Rate 1/2) transmission, respectively.

In this evaluation, 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 $m = (n + Gk) \bmod L$, $k = \{0,1,2,3\}$, $G = \lfloor L/4 \rfloor$ and $0 \leq n < (L - 1)$. That is, the LP-SS are m-sequences.

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 16 below.

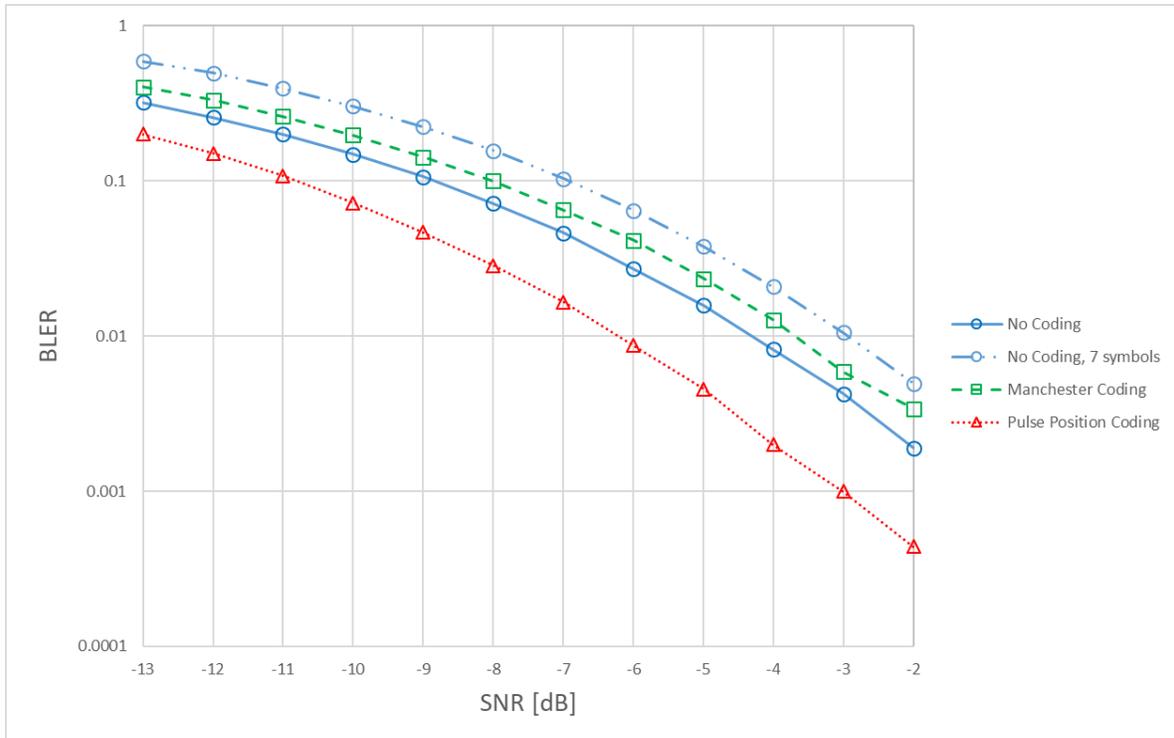


Figure 16: Detection performance for coded and uncoded LP-SS transmission, TDL-C, $M=4$.

It can be observed that a uncoded longer binary sequence ('No Coding') performs about 1.5dB better than a binary sequence of half the length ('No Coding, 7 symbols'), which is due to the improved cross-correlation among the 4 LP-SS.

Manchester coding reduces the minimum distance (or cross-correlation) between the sequences but does so under a fixed mapping which limits the degrees of freedom. The gain compared to the same uncoded sequences ('No Coding, 7 symbols') is about 0.8dB.

On the other hand, longer sequences, without the constraint of Manchester coding, can achieve lower cross-correlation and the simulation suggests a gain of about 1.4dB and 0.6dB compared to shorter uncoded sequences and Manchester coded sequences, respectively.

Observation 8: Manchester Coding for LP-SS imposes significant constraints resulting performance degradation compared to longer uncoded sequences.

Applying PPC results in a *gain* of about 3.3dB, 2.5dB and about 1.9dB compared uncoded sequences of the same length, Manchester coded sequences and to the longer sequences ('No Coding'). 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. However, power-pooling across OFDM symbols is not possible.

A further advantage of encoding the binary LP-SS is the improved interference mitigation by taking the difference of the amplitudes of adjacent OOK symbols within a codework, see Section 4.2.1. Moreover,

encoding ensures that there is always one ON-symbol per OFDM symbol and there is no instance where LP-SS is not transmitted in an OFDM symbol.

Therefore, we suggest to use PPC for LP-SS to improve performance.

Proposal 17: Support of Pulse Position Coding for LP-SS.

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.

Proposal 4: Multiplexing NR and WUS in frequency-domain is the base line.

Proposal 5: Confirm working assumption and agree Option 1-1 for OOK-4 ($M=1$)/OOK-1.

Proposal 6: Select Alternative 1 for Zadoff-Chu time-domain signal generation.

Proposal 7: For the length of the overlaid OFDM sequence in time-domain, select Alternative 2.

Proposal 8: Consider the number of maximum candidate sequences to be a function of M . Longer sequences allow for a larger number of candidate sequences with the same cyclic shifts.

Proposal 9: Consider 16 and 8 as maximum number of candidate sequences for M of 1 and 2, respectively.

Proposal 10: Consider 4 different roots for inter-cell interference mitigation.

Proposal 11: CS(s) and/or roots(s) are derived from the WUS configuration without need for explicit signaling.

Proposal 12: Reuse ON-Sequences specified from sequences available if overlaid OFDM sequences carry information.

Observation 1: Correlation receiver achieves significant gain over energy detection.

Observation 2: For $M = 4$, Pulse Position Coding achieves significant performance gain for all receiver types.

Observation 3:

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

Observation 4: A time-domain overlay code can significantly improve performance of the overlaid OFDM sequence transmission.

Proposal 14: Specify Manchester Coding as $0 \rightarrow [1\ 0]$ and $1 \rightarrow [0\ 1]$.

Observation 5: $M = 4$ with Manchester Coding has the worst coverage compared to $M = 1, 2$.

Proposal 15: For $M = 4$, consider jointly encoding multiple bits into ON pulse position to increase SNR by 3dB.

Observation 6: 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 (~ 0.1 dB) for many system configurations.

Proposal 16: Allow configuration of *Pulse Position Coding* for $M = 4$.

Observation 7: The receiver should be aware of which LP-SS sequence(s) to expect.

Observation 8: Manchester Coding for LP-SS imposes significant constraints resulting performance degradation compared to longer uncoded sequences.

Proposal 17: Support of Pulse Position Coding for LP-SS.

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 $\mathbf{y}_l \in \mathbb{C}^N$ the discrete received signal in base-band of OFDM symbol l . With M the number of OOK symbols per OFDM symbol we have $\mathbf{y}_l = [\mathbf{y}_{l,1}, \mathbf{y}_{l,2} \dots \mathbf{y}_{l,M}]$ with $\mathbf{y}_{l,i} \in \mathbb{C}^{N_M}$ 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

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

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

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

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 \mathbf{a} is known the receiver can carry out correlations with all possible transmitted sequences $\mathbf{s}_m = [s_{m1} s_{m2} \dots s_{mL}] \in \mathbb{C}^{NL}$. 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 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 $\mathbf{y}_{l,i}$ of the i^{th} OOK symbol is correlated with \mathbf{a} and the N_p peaks of the correlator output are combined to exploit frequency-diversity in fading channels i.e.

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

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

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

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

6.2. Simulation Assumptions

Unless otherwise stated, the link-level simulation assumption in Table 5 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	3 rd 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 5: Link-level simulation assumptions.