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1 Introduction

In this contribution, we discuss the limitations of NR Short Block-Length codes (SBLC) and show the benefits of transmission of small payloads without DMRS. Moreover, we outline some areas of novel coding strategies with improved performance.

In the last meeting RAN1#124-bis, although some discussions have taken place, no agreements have been reached regarding SBLC.

Agreement:	**RAN1#123**

- For the study of channel coding for small UCI with payload size of 3~11bits, at least considering:
  - 5G RM code
- Identify the justifiable drawbacks of 5G RM code, if exists, study potential solution(s).

This agreement considers the 5G RM code as a baseline and invites proponents of different encoding schemes to justify their proposals by identifying drawbacks of the 5G RM code.

We argue that the main drawback of the 5G RM code is their subpar performance compared to non-coherent codes that do not require DMRS and are decoded via sequence correlation. The main improvements stem from the following three aspects:

1. No DMRS, i.e. DMRS resources can be used to encode payload
2. Optimized code for non-coherent detection, i.e. sequence correlation
3. Improvements of PAPR via optimized sequence design

Moreover, the agreement in RAN1#122 clearly states that 6G is required to enhance the overall coverage:

Agreement:	**RAN1#122**

On enhanced overall coverage, identify coverage target(s) considering diverse use cases and device types

Thus, enhanced coverage is one of the key objectives of 6G and also includes uplink control channels. Uplink Control Information (UCI) with small payloads of up to 11 bits are common and can carry the following information.

- HARQ acknowledgement(s)

- Scheduling request (SR)
- CSI report (MCS, PMI RI, CSI-RSRP/RSRQ/SINR)

Typically, the UE coverage is limited by the UL due to the UE transmission power requirements and the limited beamforming capabilities. To maintain the best DL adaptation, the gNB requires at least simple CSI reports, e.g. RI, simple PMI (2-4 bits) and wideband CQI feedback (4 bits) or CSI-RSRx (7 bits). Hence, minimal UCI payloads range around $B = 10$ bits.

In 5G, small block length coding is used for to encode UCI for transmission on the PUCCH for $B \leq 11$ bits. One exception is the LP-WUS which also uses small block length coding. The PUCCH has been carefully designed to cater to different requirements, e.g. latency, UCI payload size and coverage. A summary of the 5G PUCCH Formats (PF) is shown in Table 1.

<table>
<tr><th>PUCCH Format</th><th>Frequency Resources [PRB]</th><th>Time-domain Resources [symbols]</th><th>DMRS</th><th>Payload Size [bits]</th><th>Design Target and use case</th></tr>
<tr><td>0</td><td>1</td><td>1-2</td><td>No</td><td>1-2</td><td>Low latency for fast ACK/NACK feedback/SR</td></tr>
<tr><td>1</td><td>1</td><td>4-14</td><td rowspan="4">Yes</td><td>1-2</td><td>Reliable ACK/NACK feedback for poor coverage scenarios</td></tr>
<tr><td>2</td><td>1-16</td><td>1-2</td><td rowspan="3">> 2</td><td>Low latency with larger payloads (CSI report)</td></tr>
<tr><td>3</td><td>1-16</td><td rowspan="2">4-14</td><td>Largest capacity</td></tr>
<tr><td>4</td><td>1</td><td>Medium capacity, code multiplexing</td></tr>
</table>

Table 1: Summary of 5G PUCCH Formats and their characteristics.

For UCI payloads $3 \leq B \leq 11$, only PF3 is able to reliably deliver such a payload under poor UL coverage conditions, since it allows for the most resources, i.e. 14 symbols and 16 PRBs which makes 2688 resource elements. However, for 14 symbols, 2 or 4 symbols are reserved for DMRS depending on configuration, resulting in a DMRS overhead of up to 28%. The DMRS are present to enable low-complexity coherent decoding, because non-coherent decoding across possible 2^B sequences of length 2688 is too computational expensive.

The study on coverage enhancements in Rel-17 concluded that PF3 (with 11bits) is indeed a bottleneck, [4]. However, the resulting normative work only included PUCCH enhancements based on dynamic repetition and DMRS bundling [5], despite the fact that DMRS-less designs offer significantly higher performance gains.

Observation 1: Coverage enhancement is one of the key KPIs in 6G. Previous studies of coverage enhancements showed that significant performance improvements in the transmission of small UCI payloads are possible.

In this contribution, we show that DMRS-less UCI transmission schemes with small payload sizes result in a significant performance gain compared to legacy 5G schemes. For instance, in one configuration, there is a 3dB SNR gain plus a 6.6dB gain in PAPR compared to 5G PF3, while maintaining low receiver complexity.

In summary, the study of enhanced transmissions schemes for small UCI payloads is well motivated by both the coverage requirements of 6G and the fact that potential large improvements over 5G RM codes are possible.

The remainder of this contributions provides an in-depth analysis of 5G RM codes and potential enhancements.

2 Characteristics of 3GPP Short Block Length Codes

In 3GPP NR, the transmissions of small packets $B \leq 11$ usually employ sequence encoding if $B \leq 2$, e.g. in PUCCH Formats 0 and 1, or Reed-Muller (RM) coding $B > 2$, [1]. RM coding has been a part of the 3GPP specifications since 3G where the simple and efficient decoding via a Hadamard transform was very appealing. Later it was used in LTE Rel-8 (TS36.212 V8.2.0) to encode UCI carrying channel quality information (CQI). First a $(20, B)$ binary block code was specified which was extended in NR Rel-15 to a $(32, B)$ block code.

However, RM codes perform far from optimal and can be significantly improved upon. Hardware has improved tremendously in 25 years and low decoding complexity is no longer a strong argument to justify the mediocre performance of RM codes.

Alternatives exist, such as the orthogonal convolutional codes used in CDMA systems, or short block-length codes for phase-modulation (e.g. QPSK) [3]. The latter are simple binary or non-binary codes which are designed for non-coherent detection when the number of signaling dimensions do not allow for orthogonal transmission. More recently, novel strategies for short block-length transmission have been studied in the Rel-17 SI on coverage enhancements [4] in order to improve coverage of PUCCH (notably PUCCH Format 3). This so called DMRS-less designs showed significant performance gains but specification has been postponed due to lack of time and consensus.

Returning to the Short Block-Length Codes in NR. When appropriate permutations of the rows of the generator matrix of the $(32, B)$ code is applied, for $B \leq 6$, the code represents a bi-orthogonal code in 32 *real* dimensions. To see the bi-orthogonal nature, the codewords must be transformed using a Hadamard transform of order 32. For $B > 6$ the code is extended by adding cosets of the base bi-orthogonal code obtained from $B = 6$. For instance, consider $B = 7$, the code is extended by its coset obtained from the 7th column of the generator matrix. It is clearly no longer a bi-orthogonal code. A similar procedure is used for the remaining 4 codes (i.e. $B = 8,9,10,11$).

To characterize the performance of the current 3GPP short block-length code, we use the set of pair-wise correlations between the transmit vectors $\boldsymbol{x}$, namely

$$\rho(m, m') = \boldsymbol{x}_m^H \boldsymbol{x}_{m'} \quad , m \neq m'$$

where $m, m' = 0,1, \dots, 2^B - 1$ is the message index. In the case of the joint estimation-detection receiver, the probability of error of any coding scheme increases monotonically with $\rho_{\text{NC,max}} = \max_{m \neq m'} |\rho(m, m')|$.

As a comparison, in Table 2, we compute the asymptotic loss $1 - \rho_{\text{NC,max}}$ of the 9 short-block length codes used in 3GPP 5G NR compared to an orthogonal signal set. For $3 \leq B < 12$, these codes are built from a $(32, B)$ binary block code and modulated using QPSK modulation. Hence there are 16 QPSK symbols per codeword. For this evaluation, we consider that the codes are mapped to the PUCCH Format 2 with 24 dimensions, e.g. 1 PRB and 2 symbols, with 8 additional symbols for DMRS. DMRS are known components of $\boldsymbol{x}_m$ and constant for all m. These symbols introduce redundancy which is independent of the transmitted data and can firstly be used to resolve the channel uncertainty and secondly to allow for often simpler receiver structures which split the estimation and detection components.

For $B = 3{,}4$ the performance is far from an orthogonal signal set even though the number of dimensions (24) is larger than 2^B. In general, for $B < 8$, where orthogonal signal sets can be easily constructed, there are clearly potential gains. This reflects the loss in signal energy due to DMRS which is significant. For $B > 4$, the performance degrades significantly, and it should be noted that even by increasing the number of dimensions beyond 24, the performance will not improve since the rate matching procedure simply repeats the bits (symbols) across frequency and time resources. The rate-matching will be beneficial for a frequency-selective channel but not for the simpler AWGN channel model considered in this comparison. Repetition in time will increase the signal energy (since there is a peak power constraint per OFDM symbol) but it will not increase the coding gain.

B	$1-\rho_{\text{NC,max}}$	$-10\log_{10}(1-\rho_{\text{NC,max}})$ [dB]
3	0.627322	2.025
4	0.466406	3.312
5	0.466406	3.312
6	0.466406	3.312
7	0.399075	3.990
8	0.349146	4.570
9	0.349146	4.570
10	0.292893	5.333
11	0.283140	5.480

Table 2: Example of asymptotic loss of the 3GPP $(32, B)$ code compared to orthogonal set when mapped to PUCCH Format 2 with 24 dimensions.

For $B > 8$, we will discuss short block-length non-orthogonal constructions that fill the gap shown Table 2 which are based on similar non-orthogonal codes to those in [3] but for lower spectral-efficiency and small block lengths.

Many companies also observed a rank deficiency of RM code for high code rates. Table 3 below, gives a more complete picture of the asymptotic loss for various output lengths N. For high code rates, we also observe rank deficiency, i.e. infinity asymptotic loss. As expected, the asymptotic loss decreases with increasing N. The decrease is not monotonic (visible for smaller B) because a DMRS symbol is added every 3 resource elements in this example.

	N														
B	4	6	8	10	12	14	16	18	20	22	24	26	28	30	32
3	5.94	3.01	1.76	2.43	2.66	2.40	2.77	2.95	2.73	2.53	2.55	2.40	2.27	2.04	2.03
4	Inf	6.79	5.94	6.37	5.40	3.80	3.99	4.20	4.14	4.19	4.30	4.36	3.70	3.17	3.31
5		6.79	5.94	6.37	5.40	3.80	3.99	4.20	4.31	4.48	4.36	4.36	3.70	3.17	3.31
6		Inf	Inf	Inf	6.53	4.35	5.05	4.20	4.31	4.48	4.36	4.47	3.70	3.17	3.31
7			Inf	Inf	7.19	7.56	6.79	5.36	5.55	5.42	5.60	4.68	4.70	4.09	3.99

8			Inf	Inf	7.19	7.56	6.79	5.62	5.55	5.42	5.60	5.46	5.34	5.33	4.57
9				Inf	7.19	7.56	6.79	6.15	5.55	5.67	5.94	5.71	5.64	5.33	4.57
10				Inf	9.82	7.56	6.79	6.15	5.94	5.67	5.94	5.71	5.64	5.33	5.33
11					Inf	Inf	Inf	8.13	6.58	6.11	6.53	6.08	5.94	5.73	5.48

Table 3: Example of asymptotic loss of the 3GPP (N,B)) code compared to orthogonal set when mapped to PUCCH Format 2.

Observation 2: The 3GPP RM code is rank deficient for high code rates.

Observation 3: The performance of 3GPP RM codes is far from optimal and there is significant room for improvement.

Proposal 1: Study novel encoding/modulation schemes for transmission of short packages.

3 Transmission with and without DMRS

The majority of codes used in NR for small payloads (e.g. PUCCH) are based on constructions using a combination of a binary channel code (short block-length Reed-Muller or Polar code), a modulation-mapping combined occasionally with an orthogonal spreading function across multiple OFDM symbols and the insertion of DMRS known to the gNB receiver. The intent is to perform channel estimation using the DMRS to allow for quasi-coherent detection at the gNB. Exceptions are PRACH and PUCCH Format 0, since non-coherent detection is implied as no explicit transmission of DMRS is part of the waveform description. Both are examples of non-binary orthogonal transmission. Interestingly, PUCCH Format 1 with 1-bit of payload is an instance of orthogonal transmission in NR, despite the presence of DMRS in the transmitted waveform.

Channel uncertainty in NR is commonly addressed by channel estimation which firstly suffers from signal energy overhead due to use of DMRS and secondly from noise enhancement due to quasi-coherent detection using the *estimates* in the place of the true channel, which in turn induces a performance penalty.

In coverage-limited scenarios, spectral-efficiency and receiver signal-to-noise ratios are both very low. As a result, the use of DMRS inherently introduces a non-negligible amount of sub-optimality that we should strive to reduce for short block-length cases. Transmission schemes without DMRS are thus to be considered when it comes to coverage enhancement system configurations.

For short block-lengths and/or low-spectral efficiency, codes can be designed for non-coherent detection (NCD). Two classes can be considered, *orthogonal* codes and *non-orthogonal* codes. When spectral-efficiency is sufficiently low it is possible to use an orthogonal transmission and this should be the chosen method. Although there is no formal proof in the scientific literature, it is widely believed that an orthogonal transmission is optimal for cases of vanishing spectral-efficiency when there is channel uncertainty at the receiver.

When the number of dimensions is not sufficiently high to use an orthogonal signal set, some form of non-orthogonal transmission is required. The conventional approach is to use DMRS signals to estimate the channel and then generate sufficient statistics for detection of the coded bit-sequence stemming from a channel code (including any rate-matching or interleaving) under the assumption that the channel is estimated perfectly. The channel code is thus constructed assuming a coherent metric in the decision rule, typically the maximum-likelihood decision rule with perfect channel state information.

Observation 4: For short block lengths, DMRS introduce a significant amount of sub-optimality and potential novel coding strategies should aim to reduce this overhead.

4 Potential Novel Techniques

A promising technique proposed during the SI on UL coverage enhancements consists of using a product-code, where the B input bits are encoded *independently* in frequency-domain (vertical) and time-domain (horizontal). This vertical and horizontal coding (VHC) strategy allows the two components to be decoded independently which reduces the complexity in the receiver.

More precisely, separate the input bits $B = B_0 + B_1$ into B_0 and B_1 bits associated with the frequency and time dimension, respectively. This is split is *optional* but can provide two levels of error protection (e.g. higher-protection for ACK/NAK than CSI/SR in a common PUCCH transmission), where by the B_0 bits can be decoded with lower error probability than the B_1 bits.

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

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

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

The design of codes $\mathcal{F}$ and $\boldsymbol{W}$ depend on the transmission requirements and the number of available dimensions. For instance, $\mathcal{F}$ should be designed to achieve low PAPR when applied to an uplink transmission. In the extreme case, where $\mathcal{F}$ contains only a single non-zero resource element per OFDM symbol the PAPR will be 0dB. Another criterion is to limit the number of non-zero elements in $\mathcal{F}$ to reduce both decoding complexity and the effects of multi-path propagation.

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

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

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

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

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

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

where $\boldsymbol{f}_{m_0 l}$ is the sequence on symbol l corresponding to message m_0. As an example, consider $N = 4$ (e.g. QPSK), $B_1 = 8$ and $L = 7$, a good generator matrix in GF4 is given by

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

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

5 Receiver Algorithms and Complexity

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

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

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

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

5.1 Non-Coherent Detection

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

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

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

5.2 Non-Coherent Detection with Reduced Complexity

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

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

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

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

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

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

The performance the low-complexity receiver can be improved by creating a list of several hypothesis for $\hat{m}_0$. In [2], we show that with a single hypothesis, i.e. $N_{\hat{m}_0} = 1$, the reduced complexity NCD performs within a fraction of a dB compared to the full NCD and with $N_{\hat{m}_0} = 2$, the performance is identical. Table 4 provides complexity comparison in terms of complex multiplications.

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

Table 4: NCD complexity comparison in terms of complex multiplications, $L = 14$, $P = 2$, $B = 11$, $B_0 = 3$ ($M_0 = 8$), $B_1 = 8$ ($M_1 = 256$).

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

In the next section, we take a look at the Coherent Detection (CD) that is used to decode the 5G NR PUCCH Format 3.

5.3 Coherent Detection

The coherent (or quasi-coherent) detection uses DMRS to estimate the channel and uses this information to decode the UCI with significantly reduced complexity. A common CD can be divided into the following steps:

1. Channel estimation
2. Softbit computation
3. Channel decoding

Denote L_{DMRS} and L_{DATA} the number of symbols with DMRS and data, respectively.

A simple channel estimation algorithm to obtain $\widehat{\boldsymbol{H}}_{l.p}$ involves a least-squares estimation on the DMRS sub-carriers as well as averaging over first frequency and then time-domain resources. This results in the same channel estimate for all time-frequency resources. The number of complex multiplications is given by KPL_{DMRS}.

The softbits for a near-ML detection (with QPSK modulation) are given by the real and imaginary parts of the matched filter output $\overline{\boldsymbol{y}}_{l,p} = \widehat{\boldsymbol{H}}_{l.p}^{H}\boldsymbol{y}_{l,p}$ which involves KPL_{DATA} complex multiplications. Subsequently, the softbits are averaged over blocks of 32 (soft)bits to obtain 32 averaged softbits as the input to the channel decoder. We omit the averaging operation in the complexity comparison.

Therefore, steps 1 and 2 require KPL complex multiplications.

One simple approach of channel decoding involves to compute the distance between the averaged softbits and all possible codewords which involves $32M$ real multiplications or $8M$ complex multiplications (omitting the additions).

A well-known low complexity channel decoder exploits the structure of the code by utilizing fast Hadamard transforms (FHT) [6]. For $B \leq 6$ there is only a single FHT and for $B > 6$, essentially, we have $32M_s$ ($M_s = 2^{B-6}$) real multiplciations of the softbits with the masks followed by M_s 32-FHTs with interleaved input. The complexity of a n-point FHT is $n \log_2 n$ additions/subtractions and can be implemented very efficiently.

Since it is difficult to account for the FHTs in terms of multiplications (since there are only additions and subtractions) we omit its complexity in the comparison. We also omit other operations such as interleaving, absolute value computation or maximum search.

Thus, the complexity of CD is approximately KPL (if $B \leq 6$) and $KPL + 8M_s$ (for $B > 6$) complex multiplications.

Table 5 compares the complexity of the receive algorithms for PF3 against the proposed VHC scheme for various parameters.

Scheme	**Number of complex multiplications**	
	$B = 6$	$B = 11$
$K = 12$ (1 PRB)		
PF3 NCD	21504	688128
PF3 CD	336	592
VHC NCD ($K_{NZP} = 1$)	1792	57344
VHC NCD-RC ($K_{NZP} = 1, N_{\widehat{m}_0} = 1$)	448	7392
VHC NCD-RC ($K_{NZP} = 1, N_{\widehat{m}_0} = 2$)	672	14560
VHC NCD-RC ($K_{NZP} = 12, N_{\widehat{m}_0} = 2$)	3136	17024
$K = 180$ (15 PRBs)		
PF3 CD ($K = 180$)	5040	5296
VHC NCD-RC ($K_{NZP} = 1, N_{\widehat{m}_0} = 1$)	1792 ($B_0 = 6, B_1 = 0$)	4032 ($B_0 = 7, B_1 = 4$)
VHC NCD-RC ($K_{NZP} = 1, N_{\widehat{m}_0} = 2$)	1792 ($B_0 = 6, B_1 = 0$)	4480

Table 5: Complexity comparison in terms of complex multiplications, L=14, P=2.

We observe that for 1 PRB, the CD for PF3 has the lowest complexity and that a full NCD has the highest complexity as expected. The complexity of the VHC NCD-RC is competitive especially for smaller payloads since less codewords are allocated to the time-domain encoding. In the extreme case of 15 PRBs, the complexity of the NCD-RC is lower than the PF3 CD but only if there is a single non-zero sub-carrier in the frequency domain allocation.

Observation 5: The proposed transmission scheme has low complexity because detection in time and frequency domain can be efficiently separated.

6 Discussion on PAPR

Reducing the PAPR of the DMRS-less PUCCH sequence is beneficial because it enables the UE to potentially transmit at a higher power. However, the reduction in PAPR does not directly translate to increased transmit power. The real gain depends on a variety of factors such as ACLR, EVM, spectrum flatness etc. and requires a study in RAN4.

The PAPR performance in illustrated in Table 6.

Scheme	**Mean PAPR** [dB]	**1% Outage PAPR** [dB]
PF3 ($\pi/2$ BPSK, 2 DMRS)	3.28	4.81
PF3 (QPSK, 4 DMRS)	3.76	6.69
VHC ($K_{NZP} = 1$)	0.03	0.06
VHC ($K_{NZP} = 3$)	2.25	2.39
VHC ($K_{NZP} = 6$)	2.35	2.61
VHC ($K_{NZP} = 12$)	2.51	2.69

Table 6: PAPR performance for $B = 4$.

It can be observed that the proposed vertical-horizontal coding scheme (VHC) offers up to 6.6 dB reduction in 1% outage PAPR compared to PF3 with QPSK and 4 DMRS. The real gain is likely much less than that but even if it enables the UE transmit at 3dB higher power, the coverage gain will be considerable.

Observation 1: DMRS-less transmission schemes provide significant room for PAPR reduction.

7 Simulation Results

In this section, we provide link-level simulation results of the BLER for the PUCCH Format 3 with CD and NCD as well as the proposed VHC scheme. Simulation assumptions are summarized in Table 7. We evaluate three different VHC schemes, two 'low PAPR' versions which use $K_{NZP} = 1$, of which one version uses the non-coherent linear block code and the second version 'low PAPR, RM' uses the 3GPP RM code. To avoid ambiguity among the codewords, the modulated RM code uses one fixed modulation symbol which is the same for all codewords, i.e. the last row of $\boldsymbol{W}$ is constant. Both schemes divide the payload $B = 11$ into $B_0 = 3$ and $B_1 = 8$, where B_1 is encoded either with the non-coherent linear block code or with the RM code.

For comparison, we added another VHC scheme termed 'medium PAPR' where $K_{NZP} = 12$, $B_0 = 4$ and $B_1 = 7$, i.e. there are 16 quasi-orthogonal sequences which encode B_0 and a non-coherent code encoding $B_1 = 7$ bits into 28 coded bits. Compared to 'low PAPR' scheme, this scheme provides better protection of the B_1 bits at the expense of the B_0 bits.

The results are shown in Figure 1.

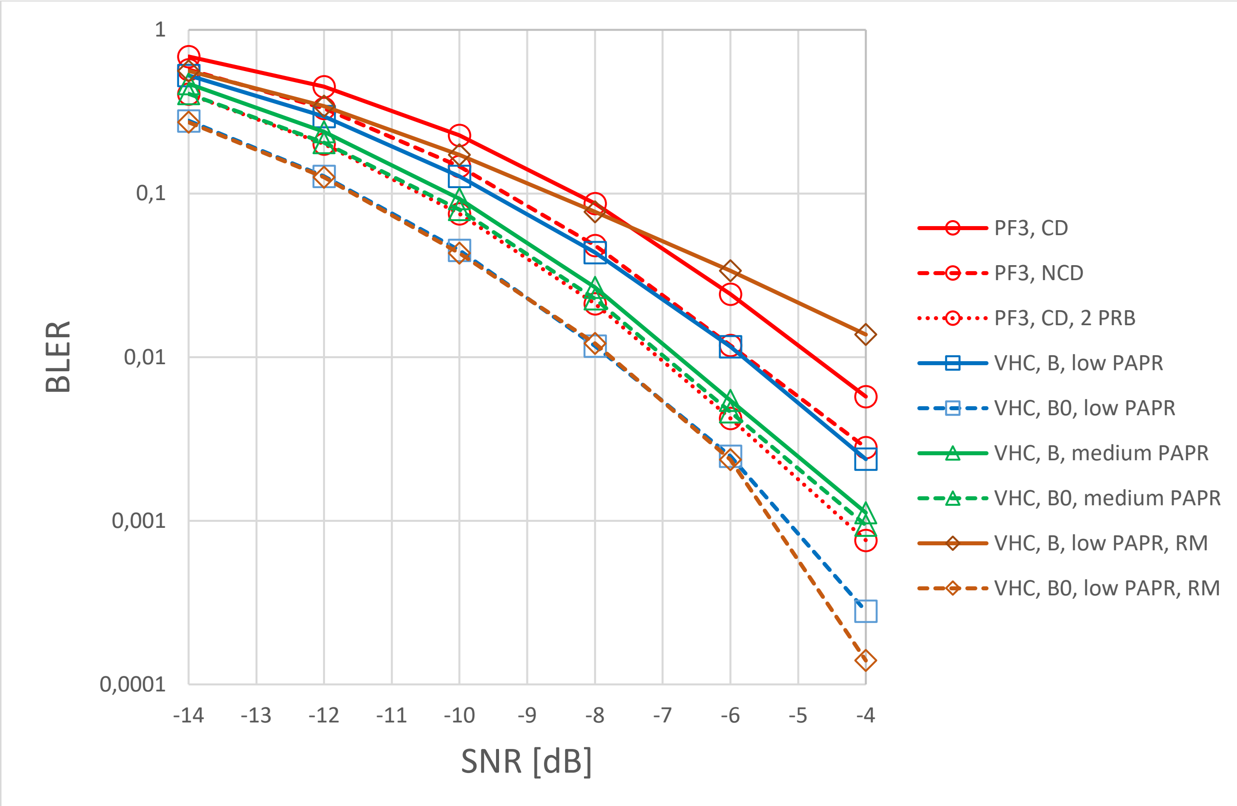

Figure 1: BLER for UCI payload of $B = 11$ bits, K =12 SCs (1 PRB) and $L = 14$ OFDM symbols.

We observe that PF3 with NCD performs about 1dB better than CD at the expense of a significantly a higher receiver complexity. Doubling the PF3 resources from 1PRB to 2PRB achieves a ~2.3 dB gain at 1%BLER.

At 1%BLER, the VHC 'low PAPR' scheme shows a 3dB gain of the B_0 bits and a 1dB gain overall compared to the standard 'PF3, CD' receiver. The performance is about the same as the 'PF3, NCD' but providing significantly lower PAPR. When using 3GPP RM coding ('low PAPR, RM') the performance of the B_0 bits is the same as the 'low PAPR' scheme as expected. However, the performance of the RM code in the encoding of the B_1 bits in time-domain results in an overall performance loss of more than 2dB, highlighting again the poor performance of this coding strategy. Lastly, the 'medium PAPR' scheme shows similar performance of the B_0 and B_1 bits, demonstrating that uneven error protection is optional and the ability to trade-off performance of the B_0 and B_1 bits.

In a second set of simulations, we compare a long PUCCH format (PF3, 1 PRB and 14 symbols) with $B = 11$ bits, 2 DMRS symbols (i.e. 144 REs for information) and 4 receive antennas in AWGN and TDL-C. No intra-frequency hopping is assumed and a coherent detection with simple LS channel estimation and averaging over 1 PRB and 14 symbols is used. It can be observed in Figure 2, that for AWGN, there is a 2.8dB gain @1%BLER when using NCD over CD for PF3. The gain of the VHC design compared to PF3, NCD is ~1dB and 1.5dB for the entire payload and the 3 bits mapped to orthogonal sequences in frequency-domain, respectively.

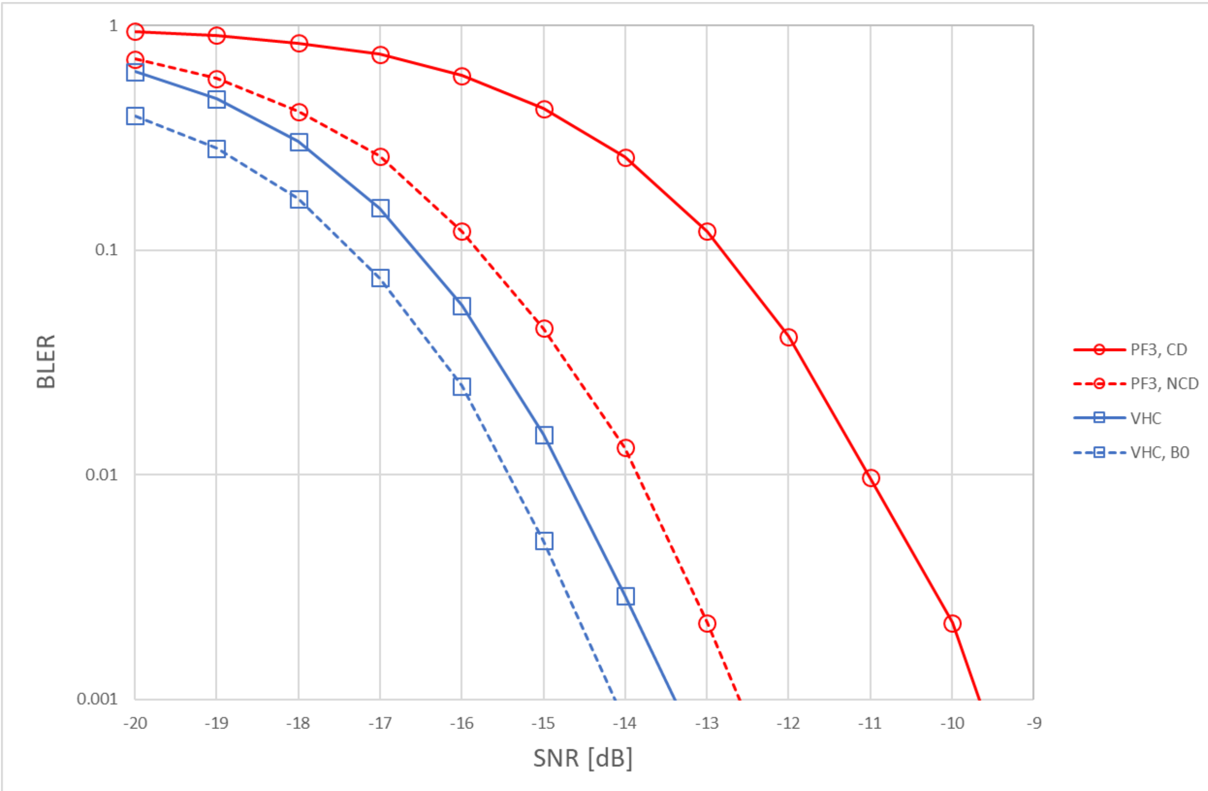

Figure 2: AWGN, 4Rx, BLER for UCI payload of B=11 bits, K=12 SCs (1 PRB) and L=14 OFDM symbols.

A similar observation can be drawn in the TDL-C channel from Figure 3, with gains of 0.6dB and 1.2dB of VHC and “VHC, B0” compared to “PF3, NCD”, respectively.

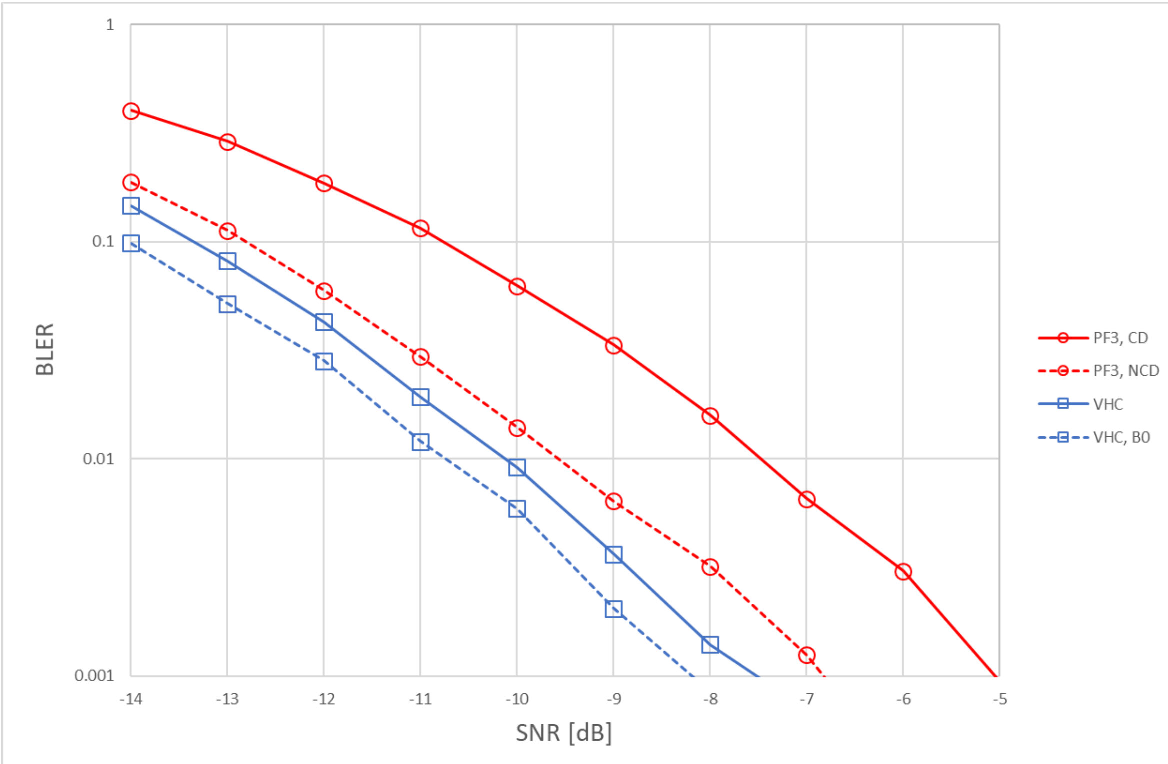

Figure 3: TDL-C, 4Rx, BLER for UCI payload of B=11 bits, K=12 SCs (1 PRB) and L=14 OFDM symbols, no hopping.

Observation 6: Simulations of novel coding strategies in UCI transmission show significant performance improvements over NR RM Codes.

8 Conclusion

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

Observation 1: Coverage enhancement is one of the key KPIs in 6G. Previous studies of coverage enhancements showed that significant performance improvements in the transmission of small UCI payloads are possible.

Observation 2: The 3GPP RM code is rank deficient for high code rates.

Observation 3: The performance of 3GPP RM codes is far from optimal and there is significant room for improvement.

Proposal 1: Study novel encoding/modulation schemes for transmission of short packages.

Observation 4: For short block lengths, DMRS introduce a significant amount of sub-optimality and potential novel coding strategies should aim to reduce this overhead.

Observation 5: The proposed transmission scheme has low complexity because detection in time and frequency domain can be efficiently separated.

Observation 1: DMRS-less transmission schemes provide significant room for PAPR reduction.

Observation 6: Simulations of novel coding strategies in UCI transmission show significant performance improvements over NR RM Codes.

9 References

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10 Appendix

Link-Level simulation assumptions are shown in Table 7 below.

Parameter	Value
Carrier Frequency	2.6 GHz (FDD)
Channel BW	100MHz (273 PRBs @ 30kHz SCS)
SCS	30 kHz
Channel Model	TDL-C, 300ns Delay Spread, 0 km/h
Number of receive antennas at gNB	2/4
Number of transmit antennas at UE	1
UCI payload size	11 bits
Frequency Hopping	Intra-slot frequency hopping enabled
PUCCH Format	PUCCH Format 3
PUCCH Resource Allocation	1 PRB with 14 symbols
Number of DMRS	4 DMRS symbols
Number of slots simulated	50,000
Receiver	Coherent and Non-coherent detection, ML
Modulation	QPSK

Table 7: Link-level simulation assumptions.