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Title:	Discussion on DMRS-less PUCCH for UL Coverage Enhancements
Document for:	Discussion and Decision

1. Introduction

Coverage enhancements have been studied in Rel-16 [1] and PUCCH Format 3 (PF3) has been identified as one of the bottlenecks. Several techniques have been proposed to increase performance of PF3, i.e. (1) DMRS-less PUCCH, (2) PUSCH-repetition-Type-B like PUCCH repetition, (3) dynamic PUCCH repetition factor indication and (4) DMRS bundling across PUCCH repetitions. Although DMRS-less PUCCH provided the most promising gains, due to a lack of time and consensus, only (3) and (4) have been agreed to be specified in the Rel-17 WI [2].

During the last email discussion on Rel-18 UL enhancements [3], still no consensus was achieved on the topic of DMRS-less PUCCH. The main concern seems to be the reported gains in link-level simulations as well as the receiver complexity. Many companies reported significant gains, while others observed only minimal gains. The WID [4] includes the following potential objective:

> 6. [Study and if necessary specify following coverage enhancements for PUCCH/UCI (RAN1)]
>
> - [DMRS-less PUCCH with UCI payload size from 3 to 11 bits]
> - [DFTS-OFDM waveform for short PUCCH with UCI payload size from 3 to 11 bits]
> - [Repetition of CSI in dynamically indicated PUSCH resources]

Furthermore, the moderator proposed the following summary [3] for PUCCH coverage enhancements:

Following points need further discussion:

- Whether short study on DMRS-less PUCCH with UCI payload size from 3 to 11 bits is included, and if so whether/how any additional scope limitation is done
- Whether short study on DFTS-OFDM for short PUCCH with UCI payload size from 3 to 11 bits is included, and if so whether/how any additional scope limitation is done
- Whether short study on repetition of CSI in dynamically indicated PUSCH resources is included, and if so whether/how any additional scope limitation is done

In the following, we present our simulation results which lead us to conclude that pursuing DMRS-less PUCCH is worthwhile and can offer significant gains.

2. 3GPP Coding For Small Block Length

For $3 \leq B \leq 11$ [5] defines a $(32, B)$ binary block code (Reed-Muller code) which, for PF3, is subsequently modulated with either a QPSK or $\pi/2$-BPSK mapping. For $B \leq 6$, the code represents a bi-orthogonal code in 32 *real* dimensions. For $B \geq 7$ the code is extended by adding cosets of the base bi-orthogonal code obtained from $B = 6$ and is no longer bi-orthogonal. The rate-matching procedure simply repeats the bits to fit the available time-frequency resources which is beneficial in frequency-selective channels but not for the simpler AWGN channel model considered in this comparison. Repetition in time will increase the signal every but will not increase the coding gain.

Table 1 presents the loss factor $1 - \rho_{NC,max}$ which is the loss of effective signal energy compared to a orthogonal signal set (i.e. where $\rho_{NC,max} = 0$) for the joint-estimation detection receiver (non-coherent receiver). For this comparison, we map the code onto PUCCH Format 2 with QPSK, 1 PRB and 2 OFDM symbols. Hence, there are 16 QPSK symbols and 8 DMRS.

From Table, 1 it can be observed that the performance loss is significant. Even 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. This reflects the loss in signal energy due to DMRS which is significant. Note that the loss is even more considerable if you use the metric for the coherent receiver. For higher number of bits, the performance degrades significantly.

B	$1-\rho_{NC,max}$	$10\log_{10}(1-\rho_{NC,max})$ [dB]
3	0.65641	1.828262
4	0.58333	2.340832
5	0.52860	2.768766
6	0.52860	2.768766
7	0.47295	3.251814
8	0.44098	3.555781
9	0.44098	3.555781
10	0.44098	3.555781
11	0.32815	4.839340

Table 1: 3GPP Short Block Length Asymptotic Performance mapped to PUCCH Format 2 with QPSK, 1 PRB and 2 symbols.

Observation 1: The asymptotic performance of 3GPP Short Block Length codes is far from optimal and there is significant room for improvement.

On the other hand, the strong structure of the codes allows for efficient low-complexity decoding via fast Hadamard transforms [6], which was surely one of the reasons they were chosen for LTE more than a decade ago. However, nowadays computational performance at the base-station has significantly improved which allows for more complex receivers if justified by the performance gain of the associated channel codes.

Observation 2: At the base-station, more complex receivers are possible if justified by considerable performance gain.

3. DMRS-less PUCCH Design for 3-11 Bit Payload

During the SI phase, efficient codes have been proposed in [7]. Essentially, we propose a product code consisting of a vertical and horizontal code which is mapped to frequency and time-domain, respectively.

More precisely, separate the number of payload bits $B = B_0 + B_1$ into B_0 and B_1 bits associated with the vertical and the horizontal dimension, respectively. Without loss of generality, the transmit message $m \in \{0, 1, \ldots, M-1\}$, $M = 2^B$, is given by $m = m_0 + m_1 M_0$ with $M_0 = 2^{B_0}$ and $M_1 = 2^{B_1}$. The transmit signal $\mathbf{R}_m^{12K\times L}$ for message m in K PRBs and L symbols is given by

$$\mathbf{R}_m = \mathbf{F}_{m_0}\mathrm{diag}(\mathbf{w}_{m_1}) \tag{1}$$

where $\mathbf{F}_{m_0}^{12K\times L} \in \mathcal{F} = \{\mathbf{F}_0, \mathbf{F}_1, \ldots, \mathbf{F}_{M_0-1}\}$ is the code sequence m_0 in the vertical domain

and $\mathbf{w}_{m_1}$ is column/codeword m_1 of the horizontal code $\mathbf{W}^{L\times M_1} = [\mathbf{w}_0, \mathbf{w}_1, \ldots, \mathbf{w}_{M_1-1}]$.

The vertical code $\mathcal{F}$ can be comprised of orthogonal low-PAPR sequences and the horizontal code $\mathbf{W}$ is either orthogonal as well (if sufficient dimensions are available) or based on a non-coherent linear block-code. An example is depicted in Figure 1, where the 8 orthogonal sequences consist of a single non-zero RE in frequency domain ($K_0 = 1$) and 2 orthogonal sequences in time-domain. Here, the sequences have 0 dB PAPR and could be employed in extreme coverage scenarios.

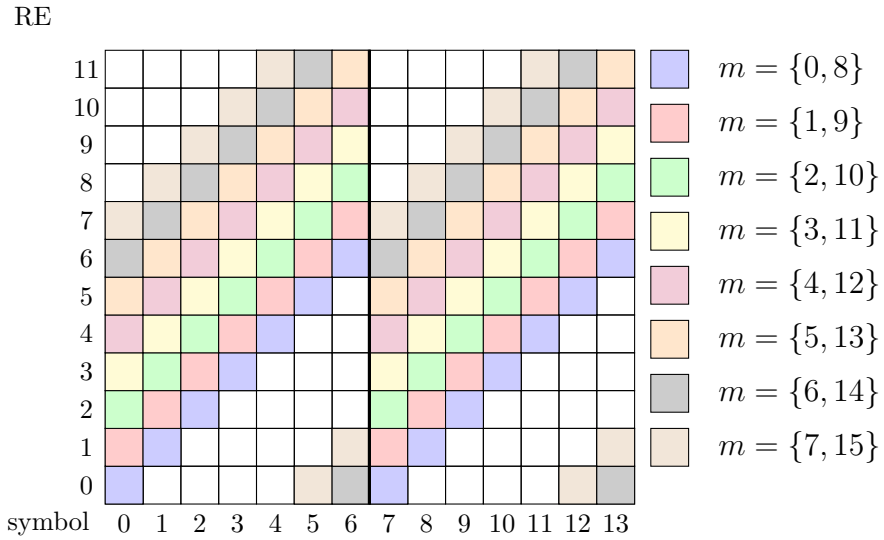

Figure 1: Example of 4 bit code structure in 1 PRB with $K_0 = 1$

In the case of an 11bit payload where $B_1 = 8$, each sequence is overlayed with $M_1 = 256$ codewords generated by a linear block code and mapped to QPSK symbols, [7].

By design, when $B_0 < B_1$, the error probability of the B_0 bits can be made lower resulting in *unequal* error protection of the payload bits. This is purely an optional feature but offers two advantages:

1. Multiplex UCI with different QoS requirements, e.g. HARQ-ACK with CSI.

2. Enable a low-complexity receiver

The low-complexity receiver is discussed in the next section.

4. Receiver Design and Complexity

In this section we discuss several receiver designs and their associated complexity.

4.1. Non-coherent Receiver

Consider the general case where the received signal $\mathbf{y}_{l,p}$ in symbol l on receive antenna $p = 1, 2, \ldots, P$ is given by

$$\mathbf{y}_{l,p} = \mathbf{H}_{l,p}\mathbf{r}_{l,m} + \mathbf{z}_{l,p} \tag{2}$$

where $\mathbf{H}_{l,p} = \mathrm{diag}(h_{l,k,p})$ with $h_{l,k,p}$ is the complex channel response on subcarrier k, $\mathbf{r}_l$ is column l of code sequence $\mathbf{R}_m$ and $\mathbf{z}_{l,p}$ is the noise term.

A near-optimal receiver of message m for a completely unknown channel realization is given by

$$\hat{m} = \arg\max_{m} \sum_{p=1}^{P} \left\| \sum_{l=1}^{L} \mathbf{y}_{l,p}^{\mathsf{H}} \mathbf{r}_{l,m} \right\|^2 . \tag{3}$$

The full non-coherent receiver in (3) has a complexity of K_0PLM complex multiplications. Recall that K_0 is the number of non-zero REs in the allocated K PRBs.

4.2. Non-coherent Receiver with Reduced Complexity

Alternatively, the detection of the vertical and horizontal code can be decoupled to allow for a lower complexity detection algorithm, which first detects the B_0 bits and subsequently uses this hypothesis to detect the remaining B_1 bits. More precisely the B_0 bits can be estimated as follows

$$\hat{m}_0 = \arg\max_{m_0} \sum_{p=1}^{P} \sum_{l=1}^{L} \left\| \mathbf{y}_{l,p}^{\mathsf{H}} \mathbf{f}_{l,m_0} \right\|^2 \tag{4}$$

where $\mathbf{f}_{l,m_0}$ is the lth column of $\mathbf{F}_{m_0}$. With estimate $\hat{m}_0$, the remaining bits B_1 associated with the horizontal code are given by

$$\hat{m}_1 = \arg\max_{m_1} \sum_{p=1}^{P} \left\| \sum_{l=1}^{L} \mathbf{y}_{l,p}^{\mathsf{H}} \mathbf{r}_{l,\hat{m}_0+m_1M_0} \right\|^2 \tag{5}$$

where $\mathbf{r}_{l,\hat{m}_0+m_1M_0}$ is the lth column of $\mathbf{R}_{\hat{m}_0+m_1M_0} = \mathbf{F}_{\hat{m}_0}\mathrm{diag}(\mathbf{w}_{m_1})$. The performance of the low-complexity receiver can be improved by creating a list of several hypothesis

in addition to the maximum in (4). Figure 2 shows the BLER performance of the full receiver in (3) compared to the low-complexity receiver with one and two hypothesis, $c = \{1, 2\}$. It can be observed that the low-complexity receiver performs within a fraction of a dB w.r.t. the full complexity receiver. With two hypothesis the performance is almost identical.

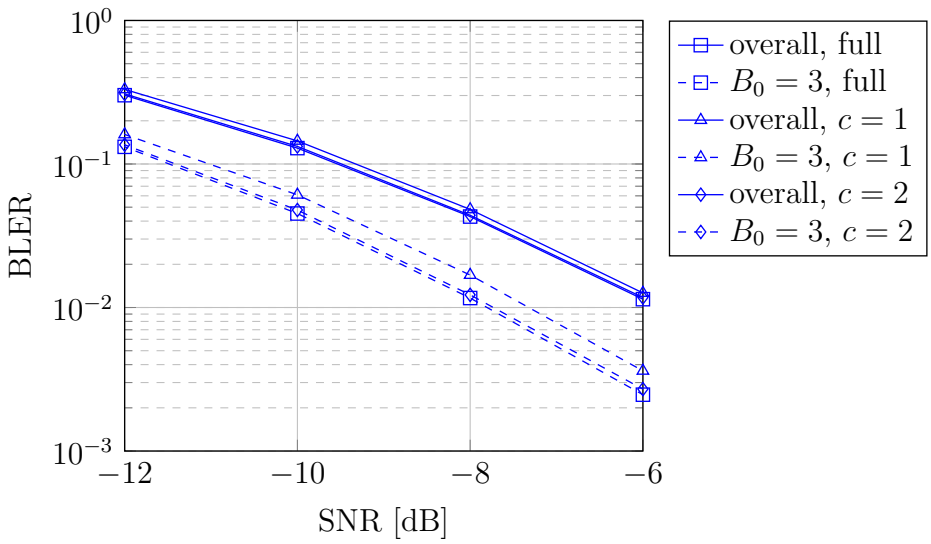

Figure 2: VHC proposal with $K_0 = 1$, 11 bits, 1 PRB.

In terms of computational complexity, the detection in (4) requires $K_0 LPM_0$ and (5) requires LPM_1 complex multiplications, respectively. Table 2 compares the number of complex multiplications for different receivers. Here, c is the number of hypothesis for the low-complexity receiver.

	Low-Complexity Rx $LPK_0M_0 + cLPM_1$								**Full Rx** LPK_0M
c	1	2	3	4	5	6	7	8	-
$K_0 = 1$	7392	14560	21728	28896	36064	43232	50400	57568	57344
$K_0 = 3$	7840	15008	22176	29344	36512	43680	50848	58016	172032

Table 2: Number of complex multiplications for $L = 14$, $P = 2$, $B_0 = 3$, $B_1 = 8$.

It can be observed that for $c = 2$ the receiver complexity is reduced by about a factor 4 and 12 for $K_0 = 1$ and $K_0 = 3$, respectively compared to the full non-coherent receiver in (3).

Observation 3: The proposed code design allows for a low complexity receiver.

4.3. Coherent Receiver

In comparison, a standard low complexity receiver for PUCCH Format 3 is a quasi-coherent detection consisting of the following stages

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 $\hat{\mathbf{H}}_{l,p}$ involves a least-squares estimation on the pilot symbols as well as averaging over first frequency and then time resources. This results in the same channel estimate for all time-frequency resources. The number of complex multiplications is given by $12KPL_{dmrs}$.

The softbits for a near-ML detection (QPSK) are given by the real and imaginary parts of the matched filter output $\bar{\mathbf{y}}_{l,p} = \hat{\mathbf{H}}_{l,p}^{\mathsf{H}}\mathbf{y}_{l,p}$ which involves $12KPL_{data}$ complex multiplications. Subsequently, the softbits are averaged over blocks of 32 bits to obtaine 32 averaged softbits as in input to the channel decoder. We omit the averaging operation in the complexity comparison.

Hence steps 1 and 2 result in $12KPL$ 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 $B < 6$ there is only one FHT and for $B > 6$, essentially, we have $32M_1$ ($M_1 = 2^{B-6}$) real multiplications of the softbits with the masks followed by M_1 32-FHTs with interleaved input. The complexity of a n-point FHT is $n\log(n)$ additions/subtractions and can be implemented very efficiently.

Since it is difficult to account for the FHTs in terms of multiplications we omit its complexity in the comparison. We also omit other operations such as interleaving, absolute value computation or maximum search.

Thus, the complexity reduces to $12KPL$ if $B < 7$ and $12KPL + M_1 8$ complex multiplications for $B \geq 7$.

Table 3 compares the receiver complexity of the different schemes in terms of complex multiplications.

Receiver	**Number Complex MUL**	
	$B = 4$	$B = 11$
VHC Low Complexity $K_0 = 1$ $c = 2$	336	14560
VHC Full $K_0 = 1$	448	57344
PF3 NCD	5376	688128
PF3 CD	336	592
PF3 CD ($K = 15$)	5040	5296

Table 3: Number of complex multiplications for $L = 14$, $P = 2$, $K = 1$.

We observe that for $B = 4$ the receiver complexity of the proposed VHC design is about the same as PF3 CD. For $B = 11$, the complexity of CD remains almost the same since it is dominated by channel estimation and matched filter computation which depend on the number of time-frequency resources. Hence, with increasing resource allocation, cf. PF3 CD ($K = 15$), the VHC design offers similar complexity than the PF3 CD.

Observation 4: The receiver complexity of a DMRS-less PUCCH design remains reasonable especially for larger number of PRBs.

5. PAPR Performance

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.

Table 4 summarizes the PAPR of the different waveforms.

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 5: DMRS-less PUCCH offers significant potential for PAPR reduction.

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 $K_0 = 1$	0.03	0.06
VHC $K_0 = 3$	2.25	2.39
VHC $K_0 = 6$	2.35	2.61
VHC $K_0 = 12$	2.51	2.69
Gold ($\pi/2$-BPSK)	3.40	4.89
Gold (QPSK)	4.27	6.51

Table 4: PAPR values for $B = 4$.

6. Simulation Results

In this section we present link-level simulation results for BLER and 4 and 11 bit payloads. The simulation configuration is summarized in Table 7. We simulate PF3 with non-coherent detection (NCD) as in (3) and with coherent detection (CD) as described in Section 4.3. The proposed VHC design uses $K_0 = 1$ and a design based on Gold sequences with $\pi/2$-BPSK mapping is also included for comparison.

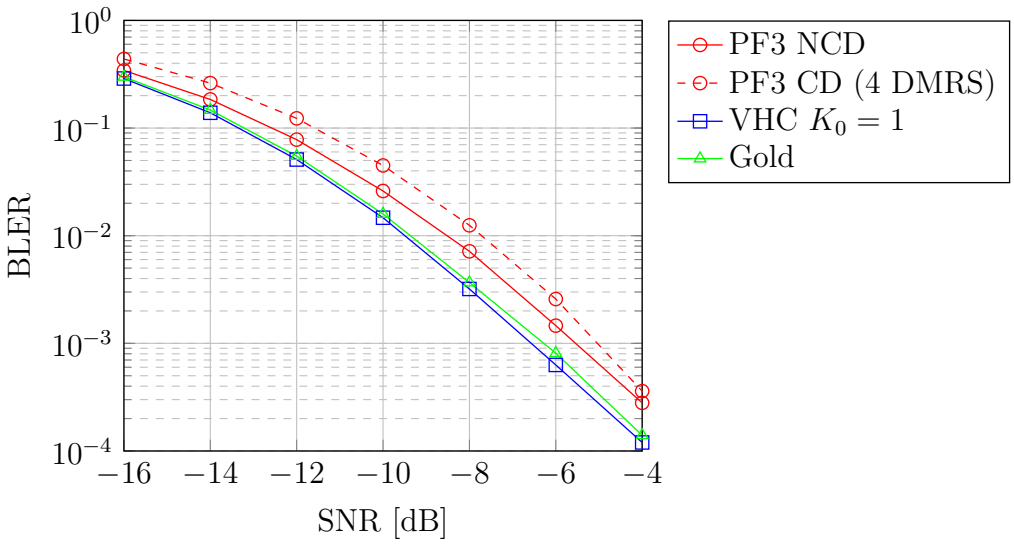

Figure 3: 4 bits, PF3 with QPSK.

From Figure 3 we observe that the VHC achieves about 1dB and 1.8dB gain @1%BLER compared to PF3 NCD and PF3 CD, respectively.

Figure 4 shows the BLER performance for 11 bits. For the VHC design we show BLER for the higher protected bits B_0, i.e. HARQ-ACK bits, and the overall BLER.

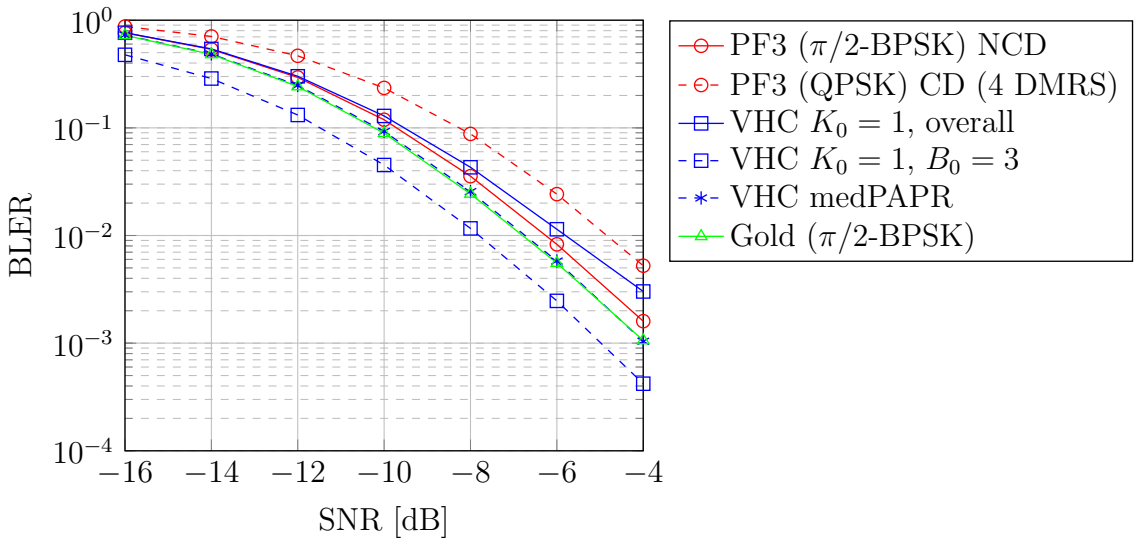

Figure 4: 11 bits, 1 PRB.

From Figure 4 we observe gains of 1.5dB and 2.9dB for the protected bits compared to the PF3 NCD and PF3 CD, respectively.

To get an idea of the gains for higher resource allocations, we simulate 3 and 15 PRBs, cf Figure 5 and 6. In the 3 PRB and 15 PRB case 5 and 7 bits are assigned to orthogonal sequences, respectively, and benefit from higher error protection.

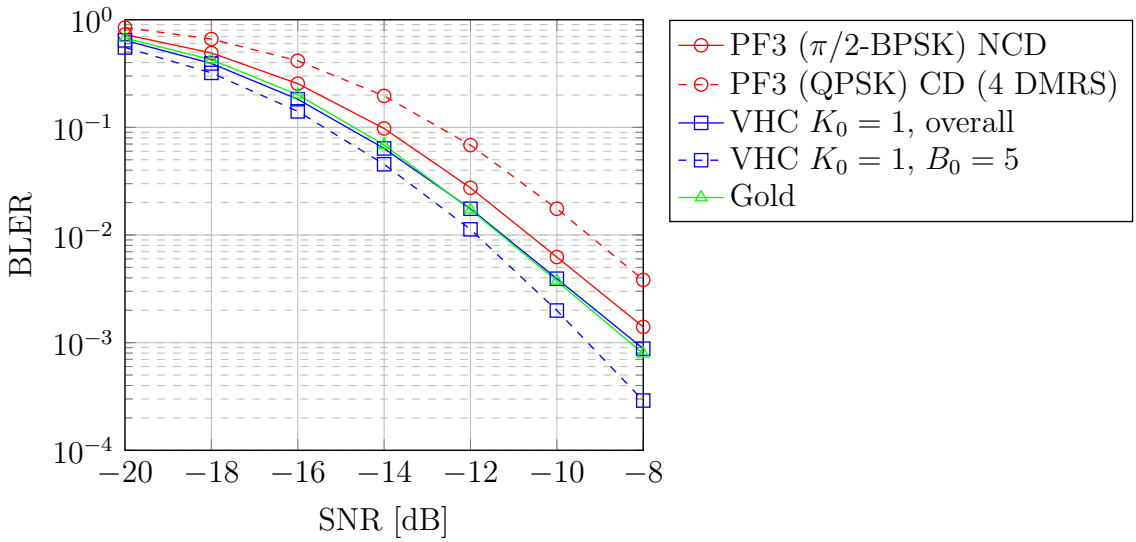

Figure 5: 11 bits, 3 PRBs.

For 3PRB, we observe a gain of 1.2dB and 2.6dB @1%BLER compared to PF3 NCD and PF3 CD, respectively. Compared to the 1 PRB case, the gain of the protected bits

has reduced but the overall gain has improved by 1dB compared to PF3 NCD.

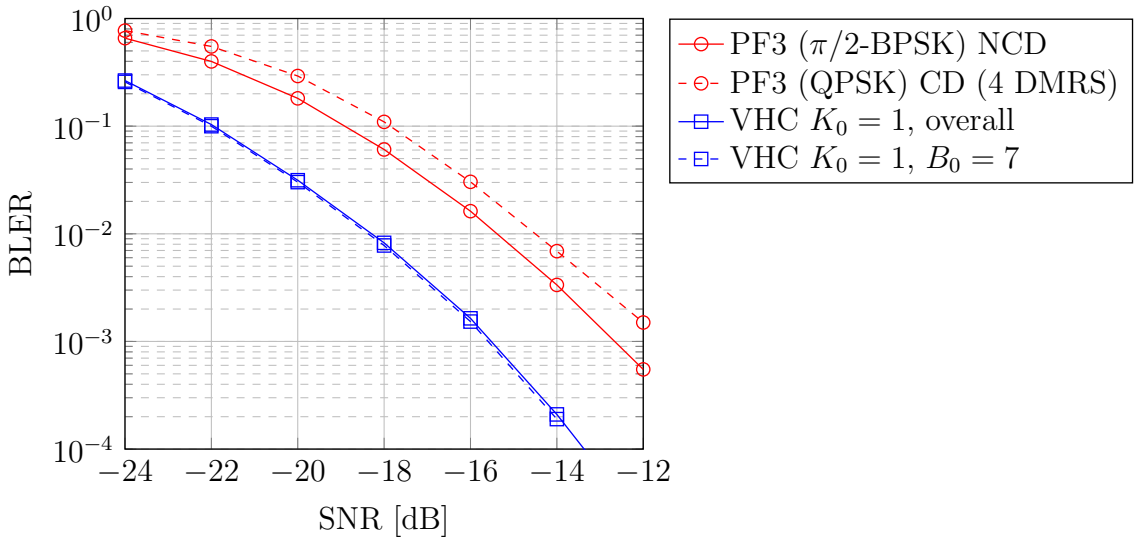

Figure 6: 11 bits, 15 PRBs.

In case of the maximum PRB allocation for PF3, i.e. 15 PRBs, we observe a gain of 3dB and 3.9 dB @1%BLER compared to PF3 NCD and PF3 CD, respectively. This result suggests that for increasing PRB allocation the gain compared to PF3 NCD and CD improves considerably.

Observation 6: The proposed DMRS-less PUCCH design offer significant BLER performance gain compared to PF3.

Observation 7: The BLER performance gain of the proposed DMRS-less PUCCH design increases relative to PF3 for increasing PRB allocation.

7. Performance Comparison

In this section we compare performance and complexity of the different schemes. The gain in Tables 5 and 6 are with respect to the VHC design overall performance. Moreover, we include the mean PAPR as opposed to the 1% outage PAPR to have a more conservative estimate for potential gain.

For a 4 bit payload, we observe from Table 5, that the proposed DMRS-less design achieves a significant SNR gain over PF3 while maintaining a similar complexity than the standard coherent PF3 receiver.

Scheme	SNR [dB] @10^{-2}	Mean PAPR [dB]	RX Complexity	Gain [dB]
VHC	-9.5	0.03	336	-
PF3 CD	-7.7	3.76	336	5.53
PF3 NCD	-8.5	3.28	5376	4.25
Gold	-9.4	3.40	5376	3.47

Table 5: Comparison for 4 bit payload.

Table 6 summarizes the gain for 11 bits. The overall gain of about 5dB compared to PF CD is considerable. An additional 2dB SNR gain is achieved for the protected bits. Interestingly, the detection of the 3 protected bits has lower complexity than the PF3 CD, the main complexity in the VHC design stems from the detection of the 8 remaining bits. Nevertheless, the overall complexity remains low.

Scheme	SNR [dB] @10^{-2}	Mean PAPR [dB]	RX Complexity	Gain [dB]
VHC $B_0 = 3$	-7.8	0.03	224	-
VHC overall	-5.8	0.03	14,560	-
PF3 CD	-4.9	4.10	592	4.97
PF3 NCD	-6.3	3.23	688,128	2.70
Gold	-6.8	3.33	688,128	2.33

Table 6: Comparison for 11 bit payload, 1 PRB.

Observation 8: DMRS-less PUCCH offers considerable gains in both SNR and PAPR while maintaining low receiver complexity.

Given the results presented in this contribution as well as during the SI phase, we are in favor of including DMRS-less PUCCH as an objective into the WID.

Proposal 1: Include DMRS-less PUCCH for 3-11 bits into the WID.

Concerning possible scope limitations, we think that already limiting the UCI size to 3-11 bits is helpful. The first main goal should be to achieve a common understanding on potential gains vs. receiver complexity of the various schemes. Hence, a short study phase seems beneficial, where proposal are evaluated through BLER performance, PAPR and receiver complexity.

Proposal 2: Include a short study phase to achieve common understanding on potential gains (SNR and PAPR) vs. receiver complexity.

8. Conclusion

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

Observation 1: The asymptotic performance of 3GPP Short Block Length codes is far from optimal and there is significant room for improvement.

Observation 2: At the base-station, more complex receivers are possible if justified by considerable performance gain.

Observation 3: The proposed code design allows for a low complexity receiver.

Observation 4: The receiver complexity of a DMRS-less PUCCH design remains reasonable especially for larger number of PRBs.

Observation 5: DMRS-less PUCCH offers significant potential for PAPR reduction.

Observation 6: The proposed DMRS-less PUCCH design offer significant BLER performance gain compared to PF3.

Observation 7: The BLER performance gain of the proposed DMRS-less PUCCH design increases relative to PF3 for increasing PRB allocation.

Observation 8: DMRS-less PUCCH offers considerable gains in both SNR and PAPR while maintaining low receiver complexity.

Proposal 1: Include DMRS-less PUCCH for 3-11 bits into the WID.

Proposal 2: Include a short study phase to achieve common understanding on potential gains (SNR and PAPR) vs. receiver complexity.

A. Appendix

A.1. Simulation Assumptions

The agreed simulation assumption can be found in the SI [1].

Parameter	Value
Carrier Frequency	2.6 GHz
Bandwidth	100 MHz (273 PRB)
SCS	30 kHz
Channel Model	TDL-C [8, Table 7.7.2-3]
Delay Spread	300 ns
MIMO Correlation	low
UE Velocity	3 km/h
Number of TXRUs at BS	2
Number of slots	100,000
PUCCH Format	PUCCH Format 3
Payload Size	4 and 11 bits
Frequency Hopping	Intra-slot frequency hopping enabled
Number of Transmit chains	1
PUCCH Duration	14 symbols
Number of PRBs	1
Modulation	QPSK
Receiver	Non-coherent and coherent
DMRS	2 or 4

Table 7: Link-level simulation assumptions for PUCCH and FR1, TDD NLOS Urban/Rural

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