Mobile Communications Beyond 52.6GHz: Waveforms, Numerology, and Phase Noise Challenge

Toni Levanen, Oskari Tervo, Kari Pajukoski, Markku Renfors, and Mikko Valkama

Abstract—In this article, the first considerations for the fifth generation (5G) New Radio (NR) physical layer evolution to support beyond 52.6GHz communications are provided. In addition, the performance of orthogonal frequency division multiplexing (OFDM) and discrete Fourier transform (DFT) spread OFDM (DFT-s-OFDM) modulations are evaluated for beyond 52.6GHz communications with special emphasis on the phase noise (PN) induced distortion. It is observed that DFT-s-OFDM is more robust against PN under 5G NR Release 15 assumptions, namely regarding the supported phase tracking reference signal (PTRS) designs, since it enables more effective PN mitigation directly in the time domain. To further improve the PN compensation capabilities, the PTRS design for DFT-s-OFDM is revised, and for the OFDM waveform a novel block PTRS structure is introduced, providing similar link performance as DFT-s-OFDM with enhanced PTRS design. We demonstrate that the existing 5G NR Release 15 solutions can be extended to support mobile communications in 60 GHz carrier frequency with the enhanced PTRS structures. In addition, DFT-s-OFDM based downlink for user data could be considered for beyond 52.6GHz communications to further improve performance with higher order modulation and coding schemes.

Index Terms—5G New Radio, 5G NR, beyond 52.6GHz, beyond 5G, DFT-s-OFDM, numerology, OFDM, phase noise, PN, phase tracking reference signal, PTRS, physical layer, PHY, SC-FDMA, spectrum availability

I. INTRODUCTION

The frequencies beyond 52.6GHz contain very large spectrum allocations and will support many high capacity use cases as envisioned in [1], such as integrated access and backhaul, ultra-high data rate mobile broadband, device-to-device communications, and industrial internet-of-things applications. Together, these would enable completely new applications for augmented or virtual reality services, factory automation, and intelligent transport systems. Operation at the millimeter waves (mmWaves) has many differences compared to lower frequencies [2], [3]. Firstly, the severely increased path loss (PL) implies that directional antenna arrays with large number of antenna elements are needed. Beam-based operation with narrow beams results in more complex channel access mechanisms, more exhaustive beam training and refinement procedures, and mainly line-of-sight (LOS) communications. Another major factor is the decreased power efficiency of power amplifiers (PAs). PAs operating in beyond 52.6GHz typically have lower output powers and are more non-linear than PAs operating on the traditional below 3GHz frequency range [4]. This implies that higher power back-off is required which directly decreases the coverage of the system. It is well known that orthogonal frequency division multiplexing (OFDM) signal has larger peak-to-average-power ratio than discrete Fourier transform (DFT) spread OFDM (DFT-s-OFDM) [5], especially at lower modulation orders, which emphasizes the importance of supporting DFT-s-OFDM in downlink (DL) and uplink (UL) for beyond 52.6GHz communications. The DFT-s-OFDM considered here corresponds to the single-carrier frequency domain multiple access (SC-FDMA) used also in Long Term Evolution (LTE) UL [5]. We note the importance of PAs for beyond 52.6GHz communications, but the focus of this work is on phase noise (PN) and its mitigation, which is another major limiting factor of radio link performance.

Already in the Third Generation Partnership Project (3GPP) Release 15 (Rel-15) New Radio (NR) air interface [2], [3], [6], PN was considered as a part of the design due to its significant effect on mmWave carriers. More specifically, 3GPP defined the so-called phase tracking reference signals (PTRSs), which allow the receiver (RX) to estimate PN from known reference symbols and compensate it before decoding the data. However, the currently supported designs may not be enough to guarantee good performance in beyond 52.6GHz communications. This is mainly due to the fact that PN increases by 6dB for every doubling of the carrier frequency [7].

In this paper, the first considerations on the physical layer (PHY) numerology and related throughput examples for beyond 52.6GHz communications are provided and discussed in detail. In the provided electronic material, a link budget tool is provided to evaluate maximum link distances for beyond 52.6GHz communications, based on the results presented in this study. In addition, we evaluate the performance of OFDM and DFT-s-OFDM with and without PTRS and PN by using the Rel-15 NR specification based implementation as baseline at 60GHz carrier frequency. The industrial, scientific and medical (ISM) band located at 60GHz carrier frequency is an important and directly available band for communications while providing up to 9GHz of bandwidth in several countries, as shown in Table I. The DFT-s-OFDM is more robust against PN induced distortion and can operate while using smaller subcarrier spacing (SCS) than OFDM, especially when high modulation orders are used. This is due to the time

T. Levanen, M. Renfors, and M. Valkama are with Department of Electrical Engineering, Tampere University, Finland (firstname.lastname@tuni.fi)
O. Tervo and K. Pajukoski are with Nokia Bell Labs, Oulu, Finland (firstname.lastname@nokia-bell-labs.com)

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domain group-wise PTRS design used with DFT-s-OFDM, which allows to track the time varying PN realization within a DFT-s-OFDM symbol. We also show that DFT-s-OFDM link performance can be further improved by designing new PTRS configurations with only modest increase in the PTRS overhead. In addition, to improve the OFDM performance, a novel block PTRS design is considered [8], allowing us to nearly achieve the link performance of DFT-s-OFDM with enhanced PTRS configuration.

II. 5G NEW RADIO ACCESS: CURRENT TECHNOLOGY AND BEYOND 52.6 GHz OPERATION

The 5G NR was designed to support wide range of SCSs to handle different uses cases and a wide range of supported carrier frequencies. In Rel-15 NR, two frequency ranges are defined, where also different SCSs are supported [2]. In Rel-15 NR, the supported SCSs follow the scaling of 15 kHz SCS with powers of two, defined as $15 \times 2^\mu$ kHz, where $\mu \in \{0, 1, 2, 3, 4\}$, corresponding to SCSs 15/30/60/120/240 kHz, respectively. The frequency range 1 (FR1) is defined for carrier frequencies 410 MHz - 7.125 GHz and supports SCSs 15/30/60 kHz, while frequency range 2 (FR2) defined for frequency range 24.25 GHz - 52.6 GHz supports 60/120/240 kHz SCSs, where 240 kHz SCS is only allowed for the so-called synchronization signal block [6]. The 5G Rel-15 NR related main PHY parameters are summarized in Table I. It is also reminded that in 5G NR numerology, the time duration of the slot and the cyclic prefix (CP) is decreased with increasing SCS and sampling frequency.

The current Rel-15 NR specification does not provide sufficient flexibility and power efficiency for communications above 52.6 GHz carrier frequencies while achieving multi-gigabit throughputs. The first component to address is the basic PHY numerology, as shown in Table I. Following the Rel-15 NR, we assume that the supported SCSs in beyond 52.6 GHz would follow 5G NR in the scaling of 15 kHz SCS with powers of two, as shown in Table I. Similar to 5G NR, we assume that FFT size of 4096 samples is used as a baseline, and that the maximum number of physical resource blocks (PRBs) equals 264, as currently defined in [9] for FR2 with 60 kHz and 120 kHz SCSs. By allocating 180 PRBs with 960 kHz SCS, the allocation bandwidth corresponds to 2.07 GHz which is well suited to the 2.16 GHz channel bandwidth, corresponding to Institute of Electrical and Electronics Engineers (IEEE) wireless local area network (WLAN) 802.11ay channel spacing [10]. The achievable PHY throughput with maximum allocation bandwidth and different modulation orders is shown in Table I. The throughput is obtained by considering a rank-1 transmission with a slot of 14 OFDM symbols, from which one symbol is reserved for physical downlink control channel (PDCCH) and one for demodulation reference signal (DMRS) used for channel estimation. In addition, PTRS overhead of 48 subcarriers corresponding to approximately 1.5% was assumed. This example shows that to reach larger than 10 Gb/s throughput, at least 2 GHz of contiguous channel bandwidth per operator should be considered.

In beyond 52.6 GHz communications, higher SCSs may be required mainly due to increased PN distortion. As the PN estimation and compensation based on PTRS is one of the main themes of this paper, it is thoroughly discussed in Section III together with considered PN models. Another reason for increased SCS is the capability to achieve extremely high bandwidth with reasonable FFT size. For reference, in IEEE WLAN 802.11ay [10], FFT size of 512 together with SCS of 5.15625 MHz is used to achieve channel bandwidth of 2.16 GHz, which can be extended with channel bonding up to maximum contiguous bandwidth of 8.64 GHz. Increasing the SCS leads also to reduced PHY latency, but it may also lead to some system design difficulties.

Use of lower SCS is desirable for several reasons. Firstly, it allows supporting longer CP length in time domain, alleviating synchronization and beam switching procedures. Secondly, it provides higher power spectral density (PSD) for transmitted signals with equal number of subcarriers. Moreover, it decreases the sampling rates required by the UE, thus enabling reduced power consumption and higher coverage for beyond 52.6 GHz communications. Furthermore, increasing the supported channel bandwidth with increased SCSs leads to increased transmitter (TX) and RX noise power which limits the coverage of the system.

### TABLE I: Physical layer numerology for 5G Rel-15 NR and considerations for beyond 52.6 GHz communications.

| Parameter                                      | Value                      |
|-----------------------------------------------|----------------------------|
| Subcarrier spacing configuration ($\mu$)       | 0 1 2 3                    |
| SCS (15 $\times 2^\mu$ kHz)                   | 15 30 60 120               |
| Sampling frequency [MHz]                       | 61.44 122.88 245.76 491.52 |
| Slot duration [us]                             | 1000 500 250 125           |
| FFT size                                       | 4096 4096                  |
| Maximum allocation bandwidth [MHz]             | 48.6 98.28 190.08 380.16   |
| Maximum channel bandwidth [MHz]                | 50 100 200 400             |
| PHY throughput with QPSK [Gb/s]                | 0.1 0.2 0.3 0.6            |
| PHY throughput with 16-QAM [Gb/s]              | 0.2 0.3 0.6 1.2            |
| PHY throughput with 64-QAM [Gb/s]              | 0.2 0.5 0.9 1.8            |
| PHY throughput with 256-QAM [Gb/s]             | 0.3 0.6 1.2 2.4            |

Beyond 52.6 GHz only supported SCSs are shown above 52.6 GHz carrier frequencies while achieving multi-gigabit throughputs.
When considering the possible available bandwidths, one has to look at current spectrum availability between frequencies 52.6 GHz and 100 GHz, as illustrated in Table II. Regarding the unlicensed access based mobile communications, global spectrum is available at frequency range 59 GHz - 64 GHz. In the current situation, best availability of unlicensed access for mobile communications is in USA, where frequency band 57 GHz-71 GHz is providing a total of 14 GHz of bandwidth. In this case, by interpolating the results shown in Table II approximately 42 Gb/s PHY throughput could be achieved with rank-1 transmission. Thus, the 60 GHz ISM band is a directly available frequency band with a vast amount of spectrum. As there are uncertainties when the other beyond 52.6 GHz frequency bands are available for mobile communications, there is a strong motivation to study the extension of the current Rel-15 NR FR2 solutions to operate also in the 60 GHz carrier frequency. It should be highlighted, that the direct extension of current FR2 operation is not a suitable long term solution when aiming to cover a wide range of carrier frequencies beyond 52.6 GHz. At the moment, the upper limit of the 3GPP related studies is set to 114.25 GHz, but even higher carrier frequencies could be envisioned in the near future, and also therefore a study on a new common waveform for beyond 52.6 GHz communications is required. In Table II the 120 kHz SCS supported in 5G NR Rel-15 FR2 operation is highlighted in bold in the context of beyond 52.6 GHz numerologies. In addition, to directly extend the current FR2 solutions to support communications in 60 GHz carrier frequency, carrier aggregation is required to support channel bandwidths beyond 400 MHz.

When looking at frequencies above 71 GHz, we note that they are mainly reserved for fixed, point-to-point communications, except in USA where both mobile and fixed communications are allowed. Considering the regulation in Europe for frequency band 71 GHz-100 GHz, we can see that up to 18 GHz aggregated bandwidth is available for fixed communications. If this could be freed also for mobile communications, then based on Table II it could be possible to achieve throughput above 54 Gb/s through frequency aggregation. In the end, the big question for beyond 52.6 GHz communications is that how the currently licensed spectrum can be efficiently shared between operators to allow ultra-high throughput operation and not to fragment the frequency bands into too small pieces.

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Considering the link budget for a beyond 52.6 GHz communications system, an example evaluation is provided in the link budget tool available through the electronic material. There, the UL direction of communications is considered as it has been traditionally the limiting direction. However, in beyond 52.6 GHz communications the DL and the UL directions become more balanced, as the same radiated power limits constraint both ends, the BS and the UE TX, respectively. The assumed TX parameterization is such that the effective isotropic radiated power (EIRP) is limited to 40 dBm following the current regulation in Europe regarding communications on frequency band 57 GHz - 66 GHz. Based on the overview of different PA output saturation powers presented in Table IV, we evaluated the arithmetic mean of these values corresponding to 15.4 dBm. Then considering the maximum power reduction required by 16-QAM modulation, we have used the value 4.5 dB in the link budget example, as defined for DFT-s-OFDM in Table 6.2.2.3-2, leading to the PA output power of 10.9 dBm. It should be noted, that for CP-OFDM, a larger maximum power reduction is required, leading to reduced maximum PA output power and coverage. In the RX side, we have used a linearly increasing noise figure with respect to the used bandwidth to include the possible effects of TX and RX non-idealities, which tend to increase with supported bandwidth. The maximum coverage provided in the link budget is based on urban micro (UMi) LOS PL model or UMi non-line-of-sight (NLOS) PL model Table 7.4.1-1. The required signal-to-noise ratio (SNR) values are taken from Fig. 4(d), presented in Section IV. Based on these assumptions and by using the available link budget tool, we note that with rank-1 transmission the theoretical maximum link distance without including the attenuation by atmospheric gases ranges from 355 m up to 3099 m in LOS channels and from 54 m up to 197 m in NLOS channels.

### III. Phase Noise Models and PTRS Designs

#### A. Phase Noise Models for Beyond 52.6 GHz Communications

A PN model is typically defined through a PSD mask, where the noise power in a 1 Hz bandwidth at a certain frequency...
offset from the carrier frequency is defined, relative to the noise power at the carrier frequency, as shown in Fig. 1. Generally, the higher is the offset, the lower is the PSD response of the PN [14]. Since OFDM and DFT-s-OFDM use multiple orthogonal subcarriers transmitted in different center frequencies, they both are affected quite similarly under PN. More specifically, PN causes a common phase error (CPE) which affects all the subcarriers similarly [8]. This means that only a single complex value is required to compensate this term from the received signal. However, due to the relatively wide PSD response of the PN in mmWave communications, it also leaks to neighboring subcarriers, which yields the so-called inter-carrier interference (ICI). This effect can be reduced or mitigated by increasing the SCS, or by applying PTRS designs which allow the estimation and compensation of the ICI components. Therefore, SCS is an important design parameter and the higher is the assumed center frequency, the higher is the required SCS, typically.

There are different PN models defined in 3GPP [13], and PN modeling has a significant effect on the radio link performance. The 3GPP models are based on extensive literature studies on current trends on phase-locked loop (PLL) based PN models and are publicly available. In general, there are two different strategies for local oscillator (LO) based carrier frequency generation. The first strategy is based on a centralized LO where a single PLL is used for transceivers, and the second strategy is based on distributed carrier generation with one PLL per transceiver. All the evaluations in this paper are based on the first strategy, i.e., we assume that there is only one PLL shared by all the transceivers, and the evaluation of distributed carrier generation is a future research topic.

In the numerical results, we use the PN model defined in [13, Section 6.1.11], which considers complementary metal oxide semiconductor (CMOS) based design for the UE due to lower cost and power consumption, and Gallium Arsenide (GaAs) based design for the BS. The GaAs based oscillators are more expensive and not as suitable to highly integrate circuits, and therefore not as well suited for UEs as CMOS based designs. The power consumption of the UE model is set to 50 mW and for the BS it is set to 80 mW, and the loop bandwidth for the PLL-based PN models is 187 kHz for the UE model and 112 kHz for the BS model. These models support by definition the 20 dB per decade scaling of the PSD, which is used to accurately match the evaluated carrier frequency of 60 GHz. Example PSD responses for the different PN models defined in [13] are illustrated in Fig. 1. The PN model defined in [13, Section 6.1.10], is an example of a PLL designed for distributed carrier generation.

B. PTRS Designs for Rel-15 NR

1) Rel-15 NR PTRS Design for OFDM: In the case of OFDM signal, individual PTRS symbols are inserted in the frequency domain with predefined frequency gap, as illustrated in Fig. 2 (a), where the physical downlink shared channel (PDSCH) carries the user data in DL direction. Thus, the Rel-15 NR PTRS design for OFDM relies on so-called distributed PTRS design in the frequency domain, occupying individual subcarriers with predefined distance in frequency. Rel-15 NR supports inserting PTRS to every second or fourth PRB in frequency domain. Since PN varies rapidly over time, PTRSs needs to be inserted densely in time. Therefore, every \( L \)th OFDM symbol in time domain, where \( L \in \{1, 2, 4\} \), can contain a PTRS. In the numerical evaluations, we assume the maximum density for Rel-15 PTRS which leads to overhead of 4.2%. Distributed frequency-domain insertion means that only CPE can be accurately estimated and compensated for each OFDM symbol containing PTRS, which may significantly limit the performance with lower SCS or high order modulations, as will be shown in Section IV.

In the RX, after channel estimation and equalization procedures, one can calculate the rotation of each PTRS in each OFDM symbol and take the average of these to obtain CPE estimate, and finally compensate it before detection and decoding procedures. In the case of not inserting PTRS to each OFDM symbol, CPE estimates for those OFDM symbols without PTRS are obtained by interpolating the available PTRS estimates in the time domain. It should be noted that interpolating CPE estimates may increase the detection latency and buffering requirements in the RX.

2) Rel-15 NR PTRS Design for DFT-s-OFDM: In the Rel-15 NR standardization phase, two different methods to insert PTRSs for DFT-s-OFDM signal were considered: pre-DFT and post-DFT insertion. That is, inserting PTRSs either in time domain or frequency domain. The latter one basically would enable exactly the same compensation mechanisms as with OFDM. However, the former one was accepted to specifications due to its lower PAPR behaviour and better PN compensation capabilities. More specifically, reference symbols are inserted before DFT to enable sample-level time domain PN tracking.

The Rel-15 NR defines different configurations for group-based time domain PTRS, where either 2 or 4 samples per group are used, and 2, 4, or 8 groups per DFT-s-OFDM symbol are supported [15, Table 6.4.1.2.2.1]. Thus, the maximum number of supported PTRSs per DFT-s-OFDM symbol is \( 8 \times 4 = 32 \), which results in overhead of 1% per DFT-s-OFDM symbol when \( 12 \times 264 = 3168 \) subcarriers are used. This maximum configuration is used as an Rel-15 NR baseline in Section IV.

The high level concept of DFT-s-OFDM PTRS is illustrated in Fig. 2 (c) together with DFT-s-OFDM symbol wise PTRS...
PN induced frequency-domain ICI components at the receiver. The idea is to allocate a frequency contiguous block of PTRS symbols, as shown in Fig. 2 (b), which allows to estimate the time varying PN within each DFT-s-OFDM symbol. In the RX, after the frequency-domain channel equalization, the received DFT-s-OFDM signal is converted back to time domain using inverse DFT, after which the PN can be estimated from the time domain ICI and compensated before detection and decoding procedures. For example, one can calculate the mean rotation in each PTRS group and use a simple linear interpolation to get the estimated PN values between the time domain PTRS groups. Note that with DFT-s-OFDM, the PTRS design allows a computationally efficient implementation to track and compensate time-varying PN response within a DFT-s-OFDM symbol, which is not possible with the Rel-15 NR distributed PTRS design for OFDM, as defined in Section III-B1 The time varying PN response over a DFT-s-OFDM symbol is observed as ICI in the frequency domain. In the case of DFT-s-OFDM, the CPE is observed as the average rotation of the time domain DFT-s-OFDM symbol.

C. PTRS Designs for Beyond 52.6 GHz Communications

1) Block PTRS Design for OFDM: The concept of frequency domain block PTRS is introduced in [8]. The basic idea is to allocate a frequency contiguous block of PTRS symbols, as shown in Fig. 2 (b), which allows to estimate PN induced frequency-domain ICI components at the receiver. As the current Rel-15 NR specification dictates a specific frequency resolution for distributed PTRS, it is possible that with block PTRS based design one can achieve better performance with lower reference signal overhead in wide channels using fullband allocations. Typically, it is considered that the block PTRS would be allocated as a multiple of PRBs, where each PRB contains 12 subcarriers, to simplify control. This is not mandated and block PTRS can also be allocated with subcarrier resolution to maximize spectral efficiency, as long as the used block size is equal or larger than the number of unknowns in the estimation process [8]. Block PTRS is inserted to each OFDM symbol, as the time continuity of ICI components is typically not guaranteed, and thus interpolation is not possible. On the other hand, having block PTRS in each OFDM symbol supports highly efficient pipelined RX architecture.

In addition to PN induced ICI, block PTRS allows to some extent compensate also the ICI induced by time varying channel and is thus well suited also for high-mobility communications where the residual Doppler error effect might be significant. Also in low-mobility scenarios in beyond 52.6 GHz communications, as will be seen in Section VI, block PTRS allows to improve the link performance with front-loaded designs (i.e., a single DMRS in the beginning of the slot), as the time varying channel during the slot duration causes ICI which is then mitigated with the block PTRS design and related compensation algorithms. In the numerical evaluations a block PTRS of size 4 PRBs, or 48 subcarriers, is assumed leading to overhead of 1.5% when assuming 3168 active subcarriers, which is clearly less than with the Rel-15 NR PTRS design.

2) PTRS Design Enhancements for DFT-s-OFDM: For beyond 52.6 GHz communications, it is important to study whether the Rel-15 NR maximum PTRS configuration is sufficient to tackle the increasing PN in the higher frequencies, or can we obtain significant performance improvements by defining new configurations. In order to improve the PN estimation capability with DFT-s-OFDM there are basically two options: 1) increasing the number of PTRS symbols per group, and 2) increasing the number of PTRS groups within the DFT-s-OFDM symbol. Increasing the number of PTRS symbols per group basically provides averaging gain against noise and interference, and does not directly improve our capability to estimate fastly changing PN response. Therefore, our proposal for the enhanced PTRS design for DFT-s-OFDM focuses in increasing the number of PTRS groups, to allow improved PN response tracking within the DFT-s-OFDM symbol. The detailed evaluation for optimized design is outside the scope of this article, and therefore a design leading to the same overhead as the block PTRS proposed for OFDM waveform was selected. Thus, we propose a new PTRS configuration for DFT-s-OFDM using 12 PTRS groups with four PTRS symbols per group, which results in 1.5% overhead when assuming 3168 active subcarriers, corresponding to the OFDM block PTRS overhead.

Fig. 2: Illustration of Rel-15 NR PTRS structures for (a) OFDM and (c) DFT-s-OFDM. In addition, the considered novel block PTRS structure for OFDM is illustrated in (b).
IV. RADIO LINK PERFORMANCE AT 60 GHz CARRIER FREQUENCY

In this section, the performance of OFDM and DFT-s-OFDM waveforms with or without PTRS and with or without PN induced interference is evaluated over the selected SCS shown in Table I and discussed in Section II. We assume a maximum channel bandwidth of 2.16 GHz, which follows the channelization for WLAN 802.11ay operating in 60 GHz unlicensed band [10]. Therefore, the maximum allocation size is limited to 180, 90, or 45 PRBs with SCS 960 kHz, 1920 kHz, or 3840 kHz, respectively. To obtain comparable performance with smaller SCS, we have limited the maximum allocation size to 180 PRBs also with SCSs less than 960 kHz.

The evaluated PN model is described in Section III-A, that is, it follows BS and UE models for TX and RX, respectively, as defined in [13, Section 6.1.11]. The evaluations concentrate on the DL rank-1 transmission scheme with polarization specific antenna panels in BS and UE. There are 128 antenna elements organized into a 8x16 antenna array per polarization at the BS and 16 antenna elements organized into a 4x4 antenna array per polarization at the UE. The link performance is evaluated with QPSK, 16-QAM, 64-QAM, and 256-QAM modulations with fixed coding rate of $R = 2/3$. The used channel codec is a 5G Rel-15 NR compliant LDPC code. The used channel model is clustered delay line E (CDL-E) with 10 ns root-mean-squared (RMS) delay spread and Rician factor $K = 15 \text{ dB}$ [12]. In all cases, a UE mobility of 3 km/h is assumed.

A. Radio Link Performance Without PTRS

First, to illustrate the significant effect of PN in beyond 52.6 GHz communications, the performance without PTRS and with or without PN is shown in Fig. 3. From the PN-free results, shown in (a) and (b), we can observe that up to 256-QAM modulation is supported with SCS larger than 120 kHz with both waveforms. With 120 kHz SCS the assumption of a slot based transmission with only 1 DMRS symbol, the channel variation caused by Doppler during the 14 OFDM symbols long slot starts to degrade the link performance, especially with 256-QAM modulation. As will be observed later on, ICI compensation allows to alleviate this problem. When considering the effect of PN, as shown in (c) and (d), we can note that QPSK modulation can be supported without PTRS with approximately 1 dB-2 dB loss in the required
SNR. Also 16-QAM could be used without PTRS at 60 GHz carrier frequency, but with significant loss in the required SNR to achieve 10% block error rate (BLER), indicating the need for PN estimation and compensation already with low-order modulations, such as QPSK and 16-QAM. High-order modulations, 64-QAM and 256-QAM, do not work at all without PTRS under PN.

B. Radio Link Performance With PTRS

The performance of OFDM using either Rel-15 NR distributed PTRS design or block PTRS design is shown in Fig. 4(a) or (c), respectively, together with DFT-s-OFDM following either Rel-15 NR based or enhanced PTRS design in Fig. 4(b) or (d), respectively. Throughout this result section, in the case of DFT-s-OFDM, we use either the time domain CPE estimate or the interpolated PN estimate, depending on which gives the best result. With OFDM and Rel-15 PTRS designs, only CPE compensation is possible. In the case of block PTRS [8], it is assumed that a contiguous allocation of 4 PRBs from the middle of the band is used for the block PTRS and on the RX side four frequency components from both sides of the DC of the PN frequency response are estimated and used in the PN compensation. It should be noted that the block PTRS based design simultaneously estimates both, CPE and ICI components.

By first comparing Fig. 4(a) and (b) with Fig. 3(c) and (d), we can conclude that the performance of low-order modulations QPSK and 16-QAM can be clearly improved with Rel. 15 PTRS, although these are typically assumed to operate without PTRS. In addition, Fig. 4(a) shows that Rel-15 NR PTRS design for OFDM can support 64-QAM if SCS ≥ 960 kHz is used. On the other hand, as shown in Fig. 4(b), DFT-s-OFDM with Rel-15 PTRS design performs significantly better with 64-QAM, allowing to use all considered SCSs. From Fig. 4(a), we also note that OFDM with Rel-15 PTRS can support 256-QAM with SCS ≥ 1920 kHz, whereas DFT-s-OFDM with Rel-15 PTRS can support 256-QAM already with SCS 240 kHz. Nevertheless, it is clear that if Rel-15 PTRS designs are not updated for OFDM or DFT-s-OFDM, significant radio link performance degradation is expected with the largest currently supported SCS of 120 kHz. The results shown in Fig. 4(a) and (b) indicate that directly extending 5G NR Rel-15 FR2 operation to 60 GHz carrier frequency implies that 16-QAM modulation can be supported by OFDM
waveform, where as DFT-s-OFDM based downlink for user data could support up to 64-QAM modulation.

Fig. 4 (c) shows that using block PTRS with OFDM improves the performance significantly when compared to Rel-15 results shown in Fig. 4 (a), and even 256-QAM can be supported with all evaluated SCs. This highlights the need for block PTRS support with OFDM in beyond 52.6 GHz communications, especially if considering the extension of 5G NR Rel-15 FR2 operation to 60 GHz carrier frequency, and also the dominance of PTRS designs allowing ICI compensation over CPE compensation. Fig. 4 (d), illustrates that increasing the number of PTRS groups from 8 to 12 with DFT-s-OFDM can improve the performance significantly. More specifically, the used 12 PTRS groups allow to support 256-QAM over all evaluated SCSs with even better link performance than OFDM using block PTRS. In addition to significant improvements with high-order modulations, also with QPSK and 16-QAM modulations clear link performance improvements are observed when comparing Figures 4 (b) and (d), thus highlighting the need for updated PTRS designs also for DFT-s-OFDM. Together, DFT-s-OFDM based downlink for user data combined with new PTRS configurations would allow to support 256-QAM modulation if current Rel-15 NR FR2 operation is extended to 60 GHz carrier frequency.

V. CONCLUSIONS AND FUTURE PERSPECTIVE

In this paper, the main use cases and implementation challenges related to beyond 52.6 GHz communications were discussed, with specific emphasis on the concept numerology and comparison of OFDM and DFT-s-OFDM waveforms with or without PN induced distortion. In addition, the performance of current PTRS designs enabled by the 3GPP 5G Rel-15 NR standard were evaluated and compared to enhanced PTRS designs, which include a novel block PTRS design for OFDM waveform and improved PTRS configurations for DFT-s-OFDM waveform.

With the extensive set of performance evaluations, it was demonstrated that the DFT-s-OFDM performs significantly better than OFDM when using Rel-15 PTRS structures, since it allows to estimate the time varying PN response within the DFT-s-OFDM symbol. Furthermore, it was shown that even QPSK performance can be improved using PTRS, and that the proposed enhanced PTRS designs can support up to 256-QAM with all the evaluated SCs. Therefore, the existing Rel-15 NR FR2 solutions can be extended to 60 GHz carrier frequency if a new block PTRS design is introduced for OFDM, or DFT-s-OFDM based downlink with new PTRS configurations is introduced for user data.

In the long run, a single solution is required for mobile communications over the whole range of 52.6 GHz - 114.25 GHz, and a simple extension of 5G NR Rel-15 FR2 solutions is not plausible. Especially, if the scope of the 3GPP studies in the future will increase the maximum supported carrier frequencies even further. The new solution will most likely include new subcarrier spacings to support wider channel bandwidths with fixed FFT size, and to enable the use of higher order modulations beyond 100 GHz carrier frequencies. In addition, new single carrier waveforms to further reduce the PAPR with different modulation orders are required to improve the power efficiency and coverage over the whole frequency range.

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