Sample Clock Offset Compensation in the fifth-generation new radio Downlink

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Sample Clock Offset Compensation in the fifth-generation new radio Downlink

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Abstract. In this paper, the authors investigate the issue of sampling clock offset (SCO) in fifth generation new radio systems. Due to the imperfect SCO estimation methods, the correction methods relying on the SCO estimation will not be perfect, so the proposed method directly corrects the effect of SCO without using any kind of estimation methods. Our method can work well in physical downlink shared channel (PDSCH) and use reference signals as recommended by the 3rd generation partnership project (3GPP) standards. Through system-level simulations, the obtained results show that the throughput of the system is improved where the gain difference between the proposed method and the case without SCO compensation arises to reach 25 % over tapped-line (TDL) channel.

1. Introduction

Modern communication systems are expected to meet the growing demands to gain better properties in terms of data rate, robustness, and reliability over multipath channels. Cyclic Prefix Orthogonal Frequency Division Multiplexing (CP-OFDM) has been considered to be used as a physical layer accessing format into 4-th generation (4G) systems and 5-th generation new radio (5G-NR) [1], because of its good performance on multipath channel, high spectral efficiency, and flexibility in bandwidth allocation [2]. However, these advantages can be obtained only if the local oscillators at both transmitter and receiver sides are synchronized with each other. In general, when the frequency spacing between subcarriers is decreased, then the sensitivity to synchronization errors is increased. Based on local oscillators mismatching in the transmitter and receiver sides, there are two different kinds of mismatches: The first one is the carrier frequency offset (CFO) which is caused by the mismatching between the radio frequency (RF) local oscillators and Doppler frequency. The second one is related to the sampling clock offset (SCO), which is induced because of the mismatching between digital-to-analog/analog-to digital converter (DAC/ADC) in the receiver and the transmitter. Without using any compensation method, these errors can destroy the receiver's performance, so we can notice that the problem of synchronization in OFDM systems is still an important subject in many researches [3]. Since it was found that the problem of sampling mismatching is more complex compared with the case of CFO, the issue of SCO has been adopted to be discussed in this paper.

T. Pollet in [4] studied the impact of SCO in the case of OFDM systems, and he found the formula that describes the bit error rate (BER) degradation due to the clock mismatches. As a result of his work, the degradation depends on the square of both carrier index and the relative frequency offset.

There are many SCO estimation methods, but generally, most of them depend on known symbols at the receiver side, the first is based on using the cyclic prefix (CP) [5]. This method is efficient since it does not require pilot symbols, but the performance gets worse compared to other methods at low SNR and high selectivity channel, we can attribute that because of the dependency of this method on CP length and delay spread of the multipath channel. The second method is based on transmitting pilot symbols and using them to estimate the value of mismatch between local oscillators [6], this method depends on either transmitting known OFDM synchronization symbols [7] or inserting pilot symbols within the OFDM symbols [8]. This method is more accurate than the previous method, but as we know, using pilot signals can degrade the spectral efficiency of OFDM systems.

The second step after estimating SCO is to correct it. Generally, there are many methods: The first depends on performing interpolation methods to obtain new samples synchronized to the transmitter's symbol rate [10]. The problem of this method is the complexity of the resampling function. The second method is based on producing a control signal which is returned to the ADC atthe receiver side [11]. However, according to [12], this method is not possible for all ADCs.

Let's imagine what happens if the results of estimation method are incorrect. Simply we can say that the accuracy of sampling mismatch correction methods depend mainly on the accuracy of estimation methods, so our proposed solution, designing an SCO effect cancelation is considered without any estimation method, and that depends on the fact that the effects of sampling offset are inter-channel interference (ICI) and phase rotation. ICI effect will be considered as additional noise because the power of the ICI is so small as mentioned in [4, 9]. So, cancelling the effect of SCO is equivalent to correcting the phase rotation of the received constellations.

The rest of this paper is organized as follows: section 2 where we describe the system model. In section 3, we present a brief review of tapped-delay channel and sampling clock model. In section 4, we describe the effect of SCO on the demodulated signal. In section 5, we derive our proposed phase estimation and correction method. Some simulation results are given in section 6. Finally, conclusion is presented in section 7.

Notation. In the following sections, theses notations will be used: $j = \sqrt{-1}$. A notation in bold is always a vector containing a finite-length sequence of symbols. $\delta(t)$ Dirac delta function.

2. System model

In the following section, we will describe the processing step of the 5G NR for physical downlink shared channel (PDSCH): Transmitter and receiver. We mainly follow the fifth-generation cellular specifications standardized by 3GPP (The 3rd Generation Partnership Project) new radio (NR) [1]. PDSCH is used for the transmission of user data. Figure 1 illustrates the system model used in this paper.

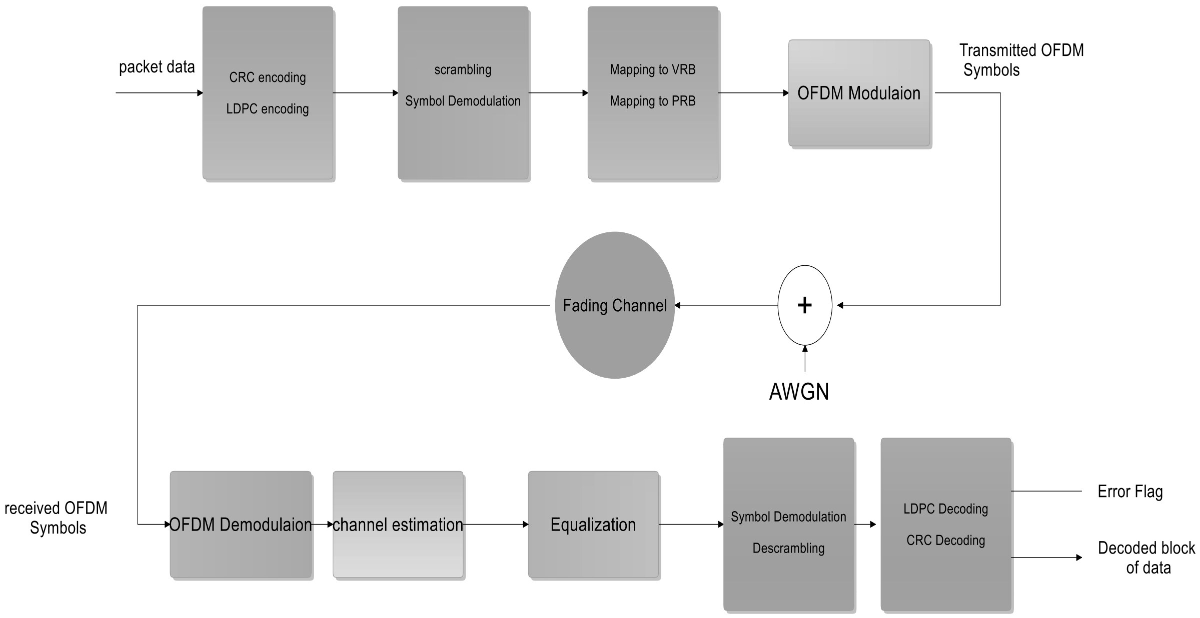

Figure 1. PDSCH physical layer System model.

The processing steps can be summarized as the following:

1) Packet data size calculation, which is based on modulation order, target code rate, modulation and coding scheme index.
2) A cyclic redundancy check (CRC) is attached to each block.
3) Each code block is encoded using low-density parity-check (LDPC).
4) Scrambling is applied to all codewords.
5) Modulation (QPSK, 16QAM, 64QAM, or 256QAM) is performed and generates complex-valued modulation symbols.
6) Resource element mapping allocates complex-valued modulation symbols to the resource elements of the resource blocks. A virtual resource block (VRB) is created for each antenna.
7) Mapping from virtual resource to PRBs is performed.
8) generating the transmitted OFDM signal.

Minimum mean square error (MMSE) equalizer per subchannel is used at the receiver side to remove the channel effects and to recover the transmitted QAM symbols. The equalized samples are then processed in the reverse order of the transmitter to give an estimate of the decoded data block and error flag signal which is the output of cyclic redundancy check (CRC) decoder.

It is worth noting that the transmissions between transmitter and receiver are based on a slot structure where every slot is defined as a period of time containing N_f OFDM symbols and N subcarriers, so an OFDM symbol can be described as: at the transmitter, N complex valued QAM symbols are modulated by N orthogonal subcarriers using an Inverse Fast Fourier Transform (IFFT). Before transmission, a cyclic prefix is appended at the start of each symbol which gives:

$$s(n) = \begin{cases} x(n+N), & if\ -N_g \le n < 0 \\ x(n), & if\ 0 \le n < N \end{cases} \tag{1}$$

where $x(n) = \frac{1}{N}\sum_{k=0}^{N-1} X_k e^{\frac{j2\pi kn}{N}}, \ 0 \le n < N$

3. Tapped-delay line channel

In this paper, tapped-delay line (TDL) channel model will be considered. In the context of signal processing, the output signal of tapped delay line filter is the sum of scaled and delayed signals from all the taps. The TDL channel can be described as a time-varying channel where the impulse response is a discrete number of impulses and can be approximated as discrete taps with different delays [13]. Channel impulse response (CIR) of TDL channel model:

$$h(\tau,t) = \sum_{l=0}^{L(t)-1} g_l(t)\delta[\tau - \tau_l(t)] \tag{2}$$

where $l = (1,2,...,L)$ — represents the index of taps, $g_l(t)$ — amplitude and phase of $l-th$ tap, $\tau_l(t)$ — delay of $l-th$ tap.

4. Frequency offset model

As we know, the real receivers mainly use local oscillators to down-convert the carrier signal to baseband, and to sample the filtered signal. To avoid synchronization errors and to decrease the bit error rate, these LOs are needed to be synchronized with LOs at the transmitter side. Figure 2 illustrates the carrier frequency offset and the sample frequency offset.

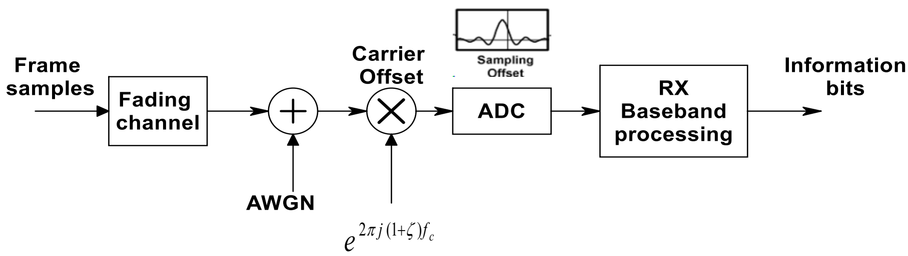

Figure 2. Modeling of carrier and sampling offset.

Modeling of carrier Frequency Offset: as we can see in Figure 2, the carrier signal at the frequency f_c is down-converted using LO with frequency equal to $(1+\zeta)f_c$. Carrier offset ζ can be described as rotation in the phase equal to $\alpha = 2\pi\zeta f_c T_s$. where T_s -the frequency sampling period. This offset will not be discussed in this paper.

Modeling of sample clock offset: If we consider that there is a sampling duration shift εT_s between sampling clock at the receiver side and sampling clock at the transmitter side, where T_s is the sampling time at the transmitter side, then the new sampling duration at the receiver side can be written as $T_s' = (1 \pm \varepsilon)T_s$, where ε is the relative sampling offset.

The received signal after removing the cyclic prefix can be written as:

$$\begin{aligned} r(n) &= \sum_{l=0}^{L(n)-1} g_l(n) s[(\varepsilon+1)n - \tau_l(n)] + w(n) \\ &= \frac{1}{N}\sum_{l=0}^{L(n)-1} g_l(n) \sum_{k=0}^{N-1} X_k e^{\frac{j2\pi k(n(\varepsilon+1)-\tau_l(n))}{N}} + w(n) \\ &= \frac{1}{N}\sum_{k=0}^{N-1} X_k e^{\frac{j2\pi kn(\varepsilon+1)}{N}} \left(\sum_{l=0}^{L(n)-1} g_l(n) e^{\frac{-j2\pi k\tau_l(n)}{N}} \right) + w(n) \\ &= \frac{1}{N}\sum_{k=0}^{N-1} X_k H_k(n) e^{\frac{j2\pi kn(\varepsilon+1)}{N}} + w(n) \\ &= y(n) + w(n) \end{aligned} \tag{3}$$

where $H_k(n) = \sum_{l=0}^{L(n)-1} g_l(n) e^{\frac{-j2\pi k\tau_l(n)}{N}}$ - the transfer function of the channel at subcarrier index k, $w(n)$ - additive white Gaussian noise (AWGN).

5. Effect of sampling clock offset

As mentioned earlier, without synchronization methods between transmitter and receiver, the performance of the system will be destroyed. To know the effect of SCO, we will process directly the received samples via the FFT without any SCO correction method, then the demodulated signal will be as follows:

$$\begin{aligned} Y_k &= \sum_{m=0}^{N-1} r(m) e^{\frac{-j2\pi km}{N}} \\ &= \sum_{m=0}^{N-1}\left(\frac{1}{N}\sum_{l=0}^{N-1} X_l H_l(m) e^{\frac{j2\pi lm(\varepsilon+1)}{N}} \right) e^{\frac{-j2\pi km}{N}} + \sum_{m=0}^{N-1} w(m) e^{\frac{-j2\pi km}{N}} \\ &= \frac{1}{N}\sum_{l=0}^{N-1} X_l \left(\sum_{m=0}^{N-1} H_l(m) e^{\frac{j2\pi m(l\varepsilon+l-k)}{N}} \right) + N_k \\ &= \frac{1}{N} X_k \left(\sum_{m=0}^{N-1} H_k(m) e^{\frac{j2\pi mk\varepsilon}{N}} \right) + \\ &= X_k \underbrace{\frac{1}{N}\left(\sum_{m=0}^{N-1} H_k(m) e^{\frac{j2\pi mk\varepsilon}{N}} \right)}_{\Sigma_k} + \underbrace{\frac{1}{N}\sum_{\substack{l=0 \\ l\neq k}}^{N-1} X_l \left(\sum_{m=0}^{N-1} H_l(m) e^{\frac{j2\pi m(l\varepsilon+l-k)}{N}} \right)}_{ICI_k} + N_k \\ &= X_k |\Sigma_k| e^{j\arg(\Sigma_k)} + ICI_k + N_k \end{aligned} \tag{4}$$

We can notice from the previous equation that the effect of SCO can be divided in three categories: phase rotation $\arg(\Sigma_k)$, amplitude reduction $|\Sigma_k|$ and inter-carrier interference (ICI).

The phase rotation also depends on channel parameters. The Channel model used in this paper is time-varying TDL model so separation between the channel phase rotation and SCO phase rotation is not possible here, and this separation can be done only if we suppose that the channel parameters do not change with time, so the previous methods that depend on measuring the value of SCO will be inaccurate in this case

The sampling offset also causes interference between carriers i.e. loss of orthogonality. Energy leakage from the ICI looks like additive noise, which in turn increases the bit error rate. In this paper, ICI is not considered so cancelling the effect of SCO is equivalent to correcting the phase rotation of constellations.

6. Proposed method

In the proposed method, we directly estimate the phase rotation caused by SCO for each symbol and compensate it.

To estimate phase rotation, a reference signal known at the receiver side is needed. The 5G NR physical layer downlink includes several reference signals such as:

- DMRS demodulation reference signals: mainly used for channel estimation
- PTRS phase tracking reference signal: used for phase tracking and noise compensation
- CSI-RS channel-state information reference signal: mainly used for channel estimation

In our method, we choose PTRS because the reference signal PTRS is allocated to a few subcarriers per symbol. It is designed to have a low density in the frequency domain and a high density in the time domain, so PTRS plays an important role in estimating the phase rotation for each symbol and minimizing the phase noise.

Since DMRS symbols are used to estimate the channel information $\hat{H}_k$, and the equalizer, in turn, uses this information to recover the transmitted symbols, then all that remains is a phase rotation which needs to be estimated and corrected.

The equalized PTRS received signal given in (4) can be written as

$$Y_k = P_k e^{j\phi_k} + Z_k \tag{5}$$

where P_k - PTRS symbols, ϕ_k is the phase rotation remained after equalizing and Z_k is the total noise, then the task will be: given P_k and Y_k for $0 \le k < N$ estimate ϕ_k .The least square(LS) estimate of the phase-rotation term is:

$$\hat{\phi}_k = \arg(\frac{Y_k}{P_k}) \tag{6}$$

Figure 3 shows the modified processing chain implemented for receiving PDSCH signal with phase rotation estimation and correction stage.

Figure 3 shows our proposed method, as we can see the proposed method runs in two modes to correct the phase rotation:

The first mode, if error flag is equal to 0, that means that the receiver has successfully decoded the block of data and the phase rotation estimation will not be used.

The second mode, if error flag is equal to 1, and that means that the receiver has failed to correctly decode the block of data, then we have to estimate the phase rotation and correct it after the equalizer.

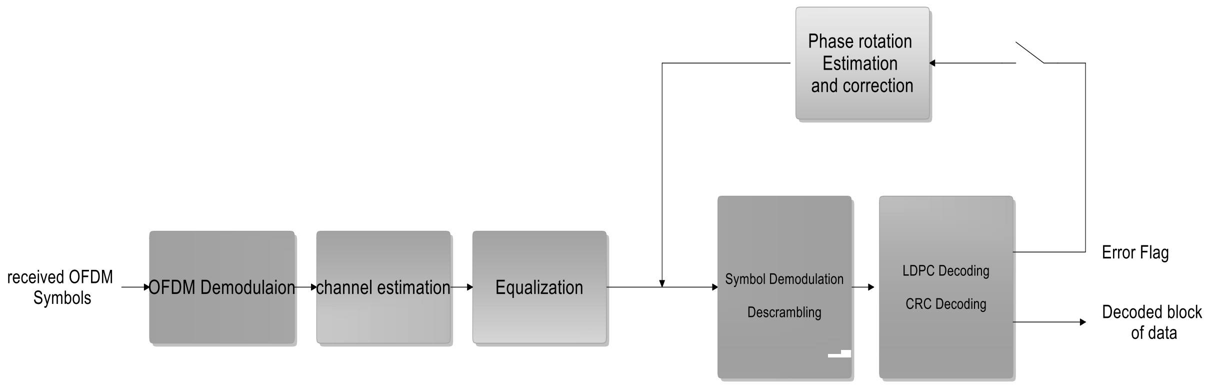

Figure 3. modified processing chain in the receiver side to compensate SCO.

7. Numerical results

Performance tests of the proposed method are carried out in MATLAB in terms of throughput as a function of SNR. The simulation parameters can be summarized as in the following table:

Table 1. Simulation parameters.

Number of frames	1000
Frame period	10 ms
Bandwidth	40 MHz
Subcarrier spacing	15 KHz
Number of useful subcarriers	300
Number of FFT points	4096
Sampling rate	61.44 MHz
Code rate	0.4785
Modulation	QPSK
Channel model	TDL

The majority of standards specify the maximum local oscillator offset to be equal to $\pm 20\ ppm$. In our simulation we will consider two values of SCO (15 ppm, 25 ppm), and show that our method works perfectly.

Figure 4 illustrates the system throughput in TDL channel and $SCO = 15\ ppm$ of the proposed method compared with system throughput without sampling offset and system throughput without using any method to correct SCO. As can be seen, there is noticeable improvement in the proposed method, where throughput gain difference between the proposed method and the case without compensating SCO in high SNR reaches 12 %.

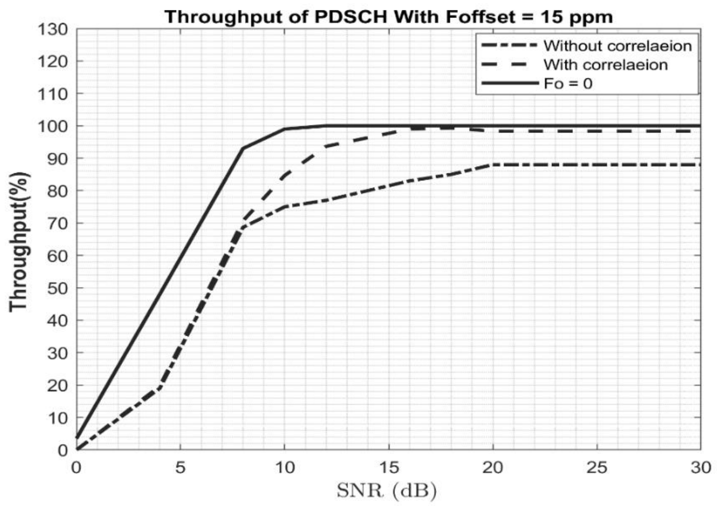

Figure 4. Throughput of the system in the case SCO = 15 ppm.

Figure 5 shows the system throughput over TDL channel and SCO = 25 ppm of the proposed method compared with system throughput without sampling offset and system throughput without using any method to correct the sampling offset. We notice that the throughput gain difference between the proposed method and the case without SCO compensation increases up to 25 %. The gap difference between throughput without SCO and proposed method is due to ICI which increases with SCO.

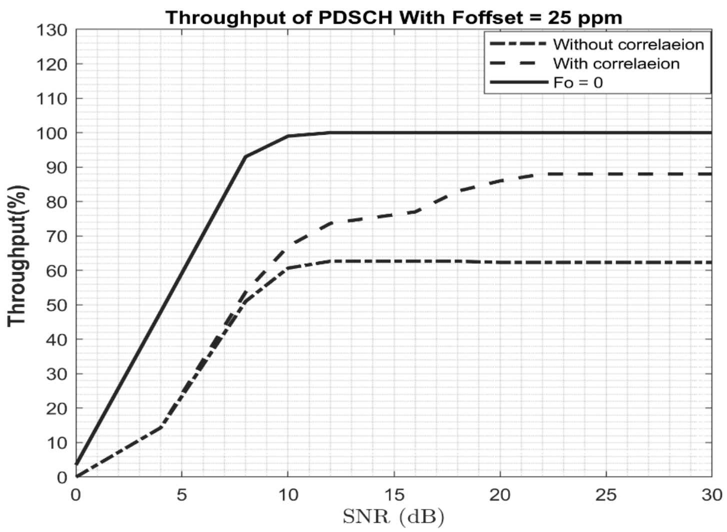

Figure 5. Throughput of the system in the case SCO = 25 ppm.

8. Conclusion

In this paper, we have addressed the issue of sampling clock offset in fifth generation new radio systems. We have reformulated all equations that are related to the effect of SCO in time- varying TDL channel and described our proposed method to compensate the phase rotation due to the SCO. We also have evaluated its performance by means of system level numerical simulations over TDL channel and different values of SCO. The obtained results show that there is a noticeable improvement in the performance when the proposed method is used. Finally, we would like to point out that for high values of SCO, the ICI will not be neglected and ICI will produce an effective noise that quickly destroy the performance, so ICI in this case have to be estimated and compensated as phase rotation.

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