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A Novel Approach for the mitigation of the ICI due to the Joint presence of Carrier Frequency Offset and Phase noise

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Abstract: In this paper an algorithm is developed for the mitigation of the ICI due to the phase noise and carrier frequency offset. A simple modification to the previously developed ICI mitigation due to the phase noise will propose a new method for the mitigation of the ICI which increase in the complexity of the receiver due to the both phase noise as well as Carrier frequency Offset. The performance of the algorithm is verified by using the DVB-T2 receiver in presence of free running Oscillator.

Index: OFDM, DVB-T2, CPE, ICI, CFO, Phase Noise and VCO.

I. INTRODUCTION

Now-a-days the entire world is converted from analog terrestrial transmission to digital terrestrial transmission because of its merits high quality of transmission of data and combat to the multi path delay. But at same time it suffers with the time varying modulation, most significantly its extreme sensitivity to the Doppler frequency shift, fast fading and oscillator jitter. The first two effects lead to a mismatch between the carrier frequencies of the received signal and the local oscillator, so that a frequency offset. Oscillator jitter also creates a very damaging effect called phase noise, meaning that the phase of the locally generated sinusoid randomly changes over time.

This paper deals with the suppression of ICI due to the presence of carrier frequency offset and phase noise. Description of the system and proposed model is given in section II while Section III proposes the implementation of a previously utilized method for mitigation of ICI due to PN alone and its subsequent modifications so that it can be efficiently applied to the more practical case of the joint presence of CFO and PN. Section IV discusses performance.

II. SYSTEM MODEL

The block diagram of the transmi- ssion is shown in figure 1 below .The proposed model is shown in figure 2.

The DVB-T2 transmitted signal of the mth symbol is given as

$$s(n) = \sum_{k=0}^{N-1} S_k e^{j2\pi kn/N}, \quad with\ n = 0,1,2,....,N-1.$$

Where $S_m(k)$ denotes the k-th subcarrier data symbol during m-th DVB-T2, symbol, interval. After the impact of a multipath channel, receiver down conversion with PN, and removal of the CP, we can write the received samples for m-th DVB-T2 symbol as a vector.

$$r(n) = s(n).e^{j\Phi(n)}$$

Here $\otimes$ circular convolution Sis the vector of the data samples.

$$R(k) = \frac{1}{N}\sum_{m=0}^{N-1} r(m) \bullet e^{-j2\pi k\frac{m}{N}}$$

$$= \frac{1}{N}\sum_{m=0}^{N-1} e^{j\Phi(m)} \sum_{r=0}^{N-1} S_r \cdot e^{\frac{j2kr\pi}{N}} e^{\frac{-j2km\pi}{N}}$$

$$= \frac{1}{N}\sum_{m=0}^{N-1} e^{j\Phi(m)} \sum_{r=0}^{N-1} S_r \cdot e^{\frac{j2\pi k(r-m)}{N}}$$

$$= \frac{1}{N}\sum_{m=0}^{N-1} S_r \sum_{r=0}^{N-1} e^{j\Phi(m)} \cdot e^{\frac{j2\pi k(r-m)}{N}}$$

In order to separate the signal and noise terms, let us suppose $\emptyset(m)$that is small, so that $e^{j\emptyset(m)}$=j$\emptyset(m)$+1.

$$R(k) \approx \frac{1}{N}\sum_{r=0}^{N-1} S_r \sum_{m=0}^{M-1} e^{\frac{j2k\pi(r-k)m}{N}}$$

$$+ \frac{j}{N}\sum_{r=0}^{N-1} S_r \sum_{m=0}^{M-1} \Phi(m).e^{\frac{j2\pi(r-k)m}{N}}$$

$$= S_k + \frac{j}{N}\sum_{r=0}^{N-1} S_r \sum_{m=0}^{N-1} \Phi(m).e^{\frac{j2\pi(r-k)m}{N}}$$

$=S_k+e_k.$

Thus we have an error term for each sub-carrier which results from some combination of all carriers and is added to the useful signal. Let us analyze more deeply this noise contribution:

1) If r=k

$$= \frac{j}{N}\sum_{r=0}^{N-1} S_r \sum_{m=0}^{N-!} \Phi(m)$$

$$= j\frac{S_k}{N}\sum_{m=0}^{N-1}\Phi(m)$$

$$= j.S_k.\Phi$$

So we have a common error added to every sub-carrier which is proportional to its value multiplied by a complex number j∅ that is a rotation of the constellation. This angle results from an average of phase noise over all of them (which imply low frequencies of phase noise spectrum) and, since it is constant for all sub-carriers, it can be corrected with the information provided by pilots.

2) If r≠k then

$$= \frac{j}{N}\sum_{\substack{r=0\\ r\neq k}}^{N-1} S_r \sum_{m=0}^{N-1}\Phi(m)e^{\frac{j2\pi(r-km)}{N}}$$

This term corresponds to the summation of the information of the sub-carriers each multiplied by some complex number which comes from an average of phase noise with a spectral shift. The result is also a complex number which is added to each sub-carrier's useful signal and has the appearance of white noise. It is normally known as inter carrier Interference (ICI) or loss of orthogonality. The ICI due to the phase noise as well as carrier frequency offset is modeled and is given as below$I_m(k)$:

$$Im(k) = \frac{1}{N}\sum_{n=0}^{N-1} e^{\frac{(2\pi fCFO\ (mN+n))}{fsamp}+\emptyset m(n))j}.$$

III. PHASE NOISE MITIGATION TECHNIQUE

This Section gives a short over view of state-of-the-art PN mitigation techniques, originally previously developed. CPE and ICI mitigation schemes are considered separately to emphasize readability.

Common phase error

As shows, CPE has exactly the same effect on every sub carrier inside one OFDM symbol. Thus, we can estimate the CPE term $J_m(0)$for an OFDM symbol by using, e.g., pre-known pilot subcarriers (S_p). To focus on CPE, we can modify so that the ICI and AWGN are just combined into one variable E_m (k).This results in

$$R_m(k) = S_m(k)H_m(k)I_m(0) + E_m(k)$$

When we consider the case kεS_p, we can estimate $I_m(O)$ with, e.g., least squares (LS) estimation, given that also the channel response $H_m(k)$ is known. This estimate can be formulated as

$$\hat{I}(0) = \frac{\sum_{k\epsilon S_p} R_m^*(k) S_m^*(k) H_m^*(k)}{\sum_{k\epsilon S_p} |S_m(k)H_m(k)|^2}$$

Where ()* is a complex conjugate operator. In, additional means to improve this estimate were also introduced for the cases where the number of pilot subcarriers (kεS_p) is low.

Inter Carrier Interference:

In CPE estimation above, only the first term of I_m vector is estimated for each OFDM symbol. All the other terms of I_m represent ICI as illustrates. The I_m vector has altogether N elements in it. We can write $R_m(k)$in more conveniently as :

$$R_m(k) = \sum_{l=-u}^{u} S_m(k-l)H_m(k-l)J_m(l) + \zeta_m(k).$$

Here, variable 'ζ_m (k) has the AWGN terms and all non-estimated ICI-terms in it.

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In practice, phase-locked-loop (PLL) based oscillators are typically used. Here, a PLL phase noise model, which contains both white and flicker noise perturbations to ¢(t),is presented. In general, the PLL PN output is dominated by the reference crystal oscillator (CO) below the loop bandwidth f_{LBw}, and by the voltage controlled oscillator (VCO) above f_{LBw}. Contemporary integrated CMOS VCO scan exhibit significant flicker noise contributions that cannot be neglected.

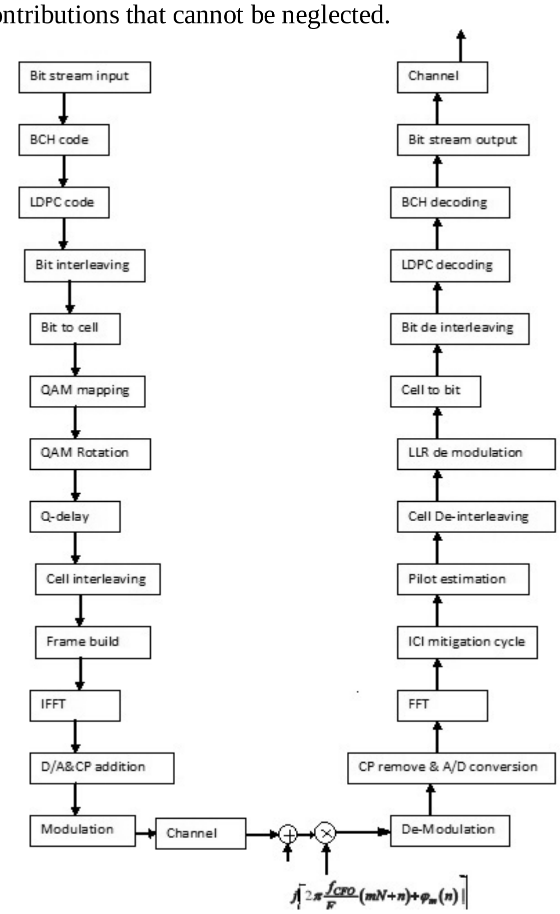

Figure 1: DVB-T2 tranmission model.

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$$\begin{bmatrix} R_m(l_1) \\ \cdot \\ \cdot \\ \cdot \\ R_m(l_p) \end{bmatrix} = \begin{bmatrix} A_m(l_1+u) & \cdots & A_m(l_1-u) \\ \cdot & \cdots & \cdot \\ \cdot & \cdots & \cdot \\ \cdot & \cdots & \cdot \\ A_m(l_p+u) & \cdots & A_m(l_p-u) \end{bmatrix} \begin{bmatrix} J_m(-u) \\ \cdot \\ \cdot \\ \cdot \\ J_m(u) \end{bmatrix} + \zeta_m$$

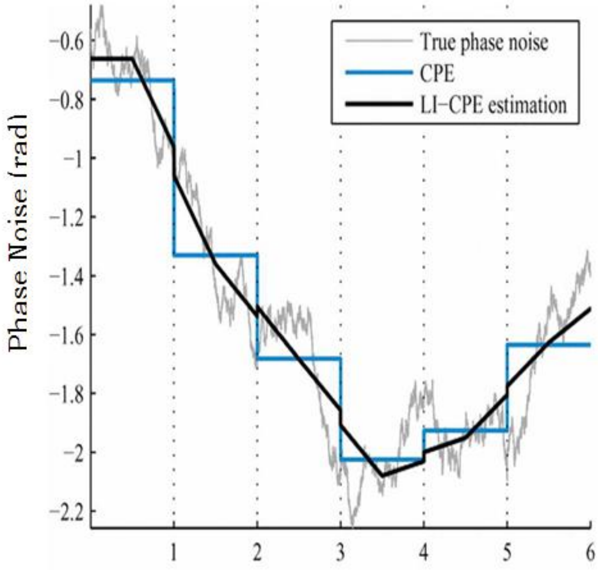
or equivalently as $R_{m,p}=A_{m,u}I_{m,u}+\tilde{\zeta}_{m,u}$, in which $A_m(k)=S_m(k)H_m(k)$.In practice, this subset of sub carriers can be selected so that it consists of sub carriers that are the most reliable after initial detection.

$$I_{m,u}=(A_{m,u}^{H}A_{m,u})^{-1}A_{m,u}^{H}R_{m,p}$$

The resulting PN spectrum estimate can then be used to deconvolve the effect of the PN out of the system.

LI-CPE estimation:

The proposed LI-CPE PN estimation technique is based on simple linear interpolation of two consecutive CPE estimates. we notice that by linearly interpolating the CPE realization from the middle of each symbol to the middle of the next symbol, our result, on average, is closer to the PN realization than the CPE estimate alone. This can be observed in the below figure: One drawback of the above estimation procedure is that it imposes an extra delay of one OFDM symbol compared to plain CPE estimation.



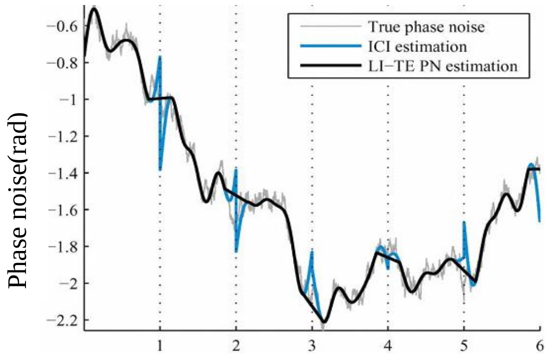
Time in OFDM symbols

Figure 2:LI-CPE methods demonst rated for a free-running oscillatorwith100 Hz spectral width over six OFDM symbols

LI-ICI estimation:

The new LI-TE PN estimation technique improves the estimation performance of iterative ICI estimation technique presented. As already noted, the ICI estimation method does not work perfectly. It has problems especially with the tails of each symbol, because Fourier series approximation does not give good PN estimates in the edges of an OFDM symbol.

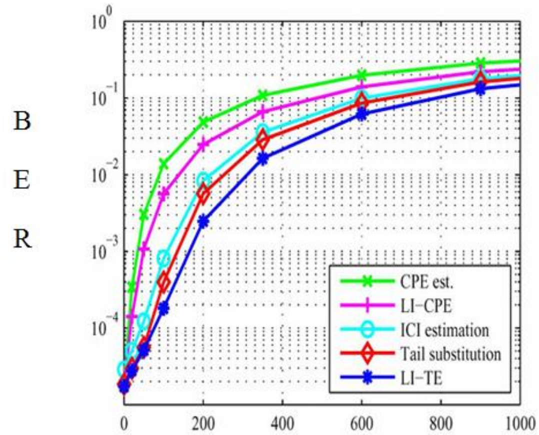
The problems can be reduced simply by linearly interpolating the phase over these badly estimated parts of the PN estimate. The linear interpolation seems to perform best when using linear interpolation over about 15%ofthe total samples from the end and the beginning of each symbol.



Time domain OFDM Symbols

Figure 3: LI-ICI methods demonst rated for a free-running oscillatorwith100 Hz spectral width over six OFDM symbols

SNR=20 dB.



SNR(dB)

Figure4. Simulated BER as a function of SNR

IV. PERFORMANCE OF THE PROPOSED MODEL

The proposed functionality is shown in figure below 5. The simulation results given below, we utilize the following DVB-T2 parameters 1K QAM constellation, normal FEC frames with 2/3 coding rate, static F1channel , c= 1/4, (PP1) [2]. The phase noise is modeled as a Wiener process (free-running oscillator) with 3dB bandwidth.

The figure 6 shows the performance of the DVB-T2 receiver in presence of the algorithm .The SNR versus BER characteristics is shown for the data, for AWGN channel, Static channel, introduction of I in AWGN, introduction of I in static channel, ICI mitigation due to phase noise in AWGN, ICI mitigation due to phase noise in static channel and finally the mitigation of the phase noise and CFO in DVB-T2 receivers.

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Both CFO and PN have been shown to be addressed in an efficient manner by the proposed method and subsequent modifications.

The computational complexity increase due to the introduction LI-ICI-E1 and LI-ICI-E2 is minimal [5] because of its location outside of the iterative chain depicted so for the reason simple ICI mitigation is used to reduce the computational efficiency and reduce the complexity of design.

As discussed above, when the proposed method is switched off, it is equivalent to setting u=0 or simple CPE correction. Additionally, for comparison with the simple CPE correction in the presence of CFO, we also explore a computationally inexpensive method [5] for estimating of CFO below ±0.5 subcarrier spacing. Where Lis the set of sub-carrier indices where pilots xexist for adjacent OFDM symbols, i.e. as with continual pilot indices of the used PP as in PP2 [2]. Its presence in the simulation results is marked by substituting 'M' with 'Pi'. We note that is suitable only for the estimation of CFO, but is inept at estimating PN or a combination of CFO and PN. Hence, this method is only given as a [5] reference.

Coded BER vs. DVB-T2 mode is shown for SNR = 16 dB with Fcfo =100 Hz,

f3dB= 0 Hz, 64-QAM, u= 4, and P= 40. With the 32K mode, the 100 Hz CFO corresponds to 36% of f= 279 Hz and is clearly a limiting factor for performance. Yet, the proposed technique provides an improvement even with low number of estimated ICI components u. A further increase in u and proportional increase of P, to maintain the system in (3) equally over determined, is shown next in Fig. 6 to improve BER in difficult conditions.

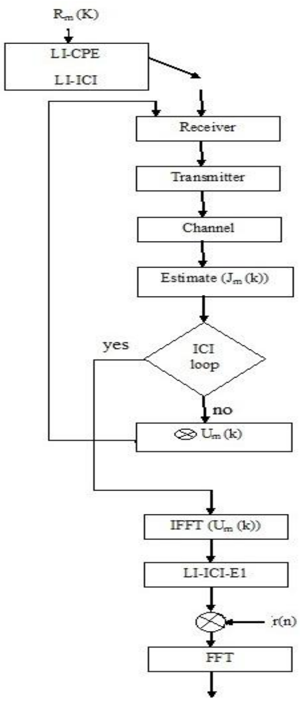

Figure 5: shows the proposed algorithm to reduce the ICI due to Phase noise and Carrier Frequency Offset.

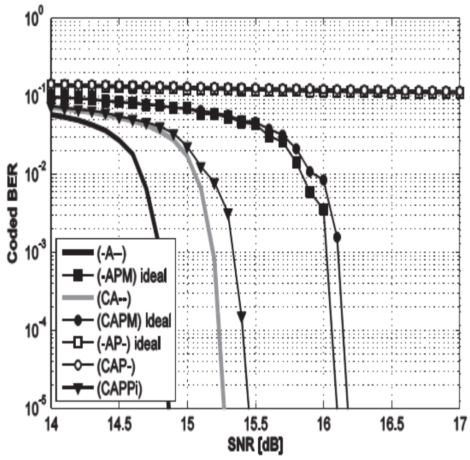

Figure 6. BER vs. SNR: 8K, 64-QAM, f_{CFO}=125 Hz.

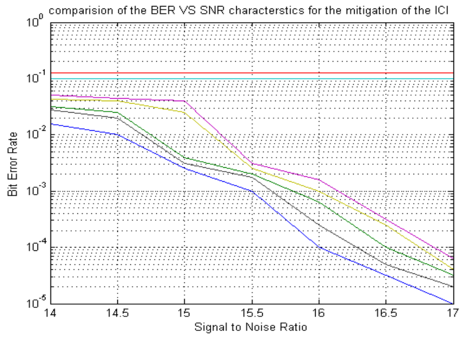

Figure 7. BER vs. SNR: 8K, 256-QAM, f_{CFO}=125 Hz.

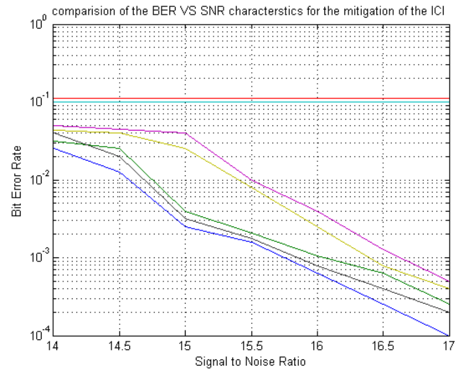

Figure 7. BER vs. SNR: 8K, 64-QAM, f_{CFO}=125 Hz.

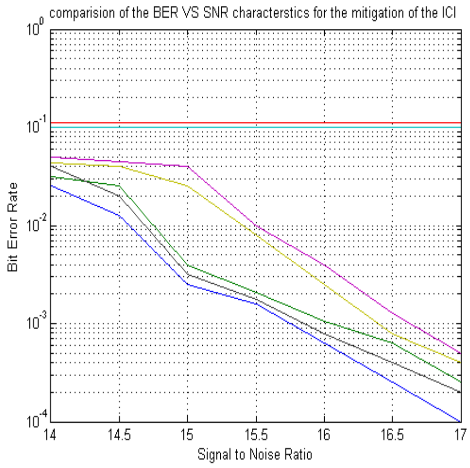

Figure.8 BER vs. SNR: 8K, 16-QAM, f_{CFO}=125 Hz.

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CONCLUSIONS

A single algorithm was shown to mitigate ICI due to the joint presence of CFO and PN. Further steps for the development and optimization of the algorithm performance are described.

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