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SPECTRUM REGROWTH REDUCTION METHOD FOR DFTS-OFDM SIGNAL

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Abstract:
One significant problem of using discrete Fourier transform spreading-OFDM (DFTS-OFDM) is to produce the undesirable spectrum regrowth at the outside of allocated bandwidth due to the non-continuity between time domain symbols. To solve this problem, this paper proposes a simple spectrum regrowth reduction method for DFTS-OFDM by inserting interpolated samples for improving the continuity between symbols. This paper demonstrates the effectiveness of proposed method as comparing with conventional spectrum regrowth reduction methods by simulation results.

Keywords: DFTS-OFDM, spectrum regrowth, interpolation, ACLR

Classification: Wireless communication technologies

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1 Introduction

Discrete Fourier transform spreading-orthogonal frequency division multiplexing (DFTS-OFDM) has been received a lot of attentions as an alternative technique to OFDM from its lower peak to averaged power ratio (PAPR) and robustness to multipath fading [1]. There is however yet remaining one significant problem of using DFTS-OFDM which produces the undesirable spectrum regrowth due to the

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non-continuity between the time domain symbols. The undesirable spectrum regrowth leads the serious interference problem to adjacent channels in the operation of DFTS-OFDMA system. To solve this problem, various methods were proposed both for OFDM and DFTS-OFDM [2-5]. The phase-anchored [2] and zero-tail [3] methods were proposed for DFTS-OFDM which can reduce the spectrum regrowth with almost the same processing load as that for DFTS-OFDM. The filtering [4] method was proposed for OFDM however it is also possible to apply to DFTS-OFDM. This method requires a small additional processing load at the transmitter with keeping almost the same processing load as the OFDM/DFTS-OFDM at the receiver. The *N*-continuous OFDM method [5] was proposed for OFDM which requires considerable additional processing load at the transmitter and especially at the receiver to achieve almost the same bit error rate (BER) performance as that for OFDM. Although above all methods can suppress the spectrum regrowth to some extent, there could be yet room for further improvement of spectrum regrowth with keeping lower processing load.

From the above background, this paper proposes a simple spectrum regrowth reduction method for DFTS-OFDM by inserting interpolated samples between the time domain symbols. This paper presents computer simulation results to demonstrate the effectiveness of proposed method as comparing with the conventional methods when applying to DFTS-OFDM.

2 Proposal of spectrum regrowth reduction method

Fig.1(a) shows the structure of proposed DFTS-OFDM transmitter. In the DFTS-OFDM, M time domain data information are converted to the frequency domain by M-points DFT and M frequency data are assigned within N subcarriers at the subcarriers mapping. After the subcarriers mapping, N subcarriers are converted to the time domain $b(i, l)$ by N-points IFFT. Then the guard interval (GI) with the length of N_g samples is added at the start of every data symbol. The time domain signal $c(i, n)$ with GI at the n-th time sample of i-th symbol is given by,

$$[c(i,n)]=[b(i,N-N_g),\cdots,b(i,N-1),b(i,0),\cdots b(i,N-N_g),\cdots b(i,N-1)],\quad 0\le n\le N+N_g-1 \qquad (1)$$

Fig. 1(b) shows the frame format and schematic diagram for proposed method. In the proposed method, Q interpolated samples are inserted between two time domain symbols so as to improve the continuity at the output of N-points IFFT. The interpolated samples are generated by using the polynomial approximation based on the minimum square error (MSE) method in which P samples both at the last and start of two symbols, and the middle point of Q interpolated samples are employed as the reference samples. From (1) and Fig. 1(b), the reference samples $y(s)$ at the last and start of $c(i$-1, $n)$ and $c(i, n)$ both with P samples and the middle point $y(m)$ are given by,

$$y(s)=\begin{cases} c(i-1,N-P-1+s), & 1\le s\le P \\ \{c(i-1,N-1)+c(i,N-N_g)\}/2, & s=m \\ c(i,N-N_g-P-Q-1+s), & P+Q+1\le s\le 2P+Q \end{cases} \qquad (2)$$

where s represents the sample number from 1 to $2P+Q$ and m is the middle point of Q samples which is given by $m=[(2P+Q)/2]$. Here [x] represents the nearest integer higher than x. The reference sample of middle point $y(m)$ is given by using the linear interpolation between the reference samples $y(P)$ and $y(P+Q+1)$ as shown in Fig.1(b). The interpolated samples Q between two symbols are approximated by the polynomial expression with L orders which is given by,

$$\hat{y}(s) = \sum_{k=0}^{L-1} \beta_k \cdot s^k, \quad 1 \le s \le 2P+Q \tag{3}$$

where β_k is the unknown coefficient for the polynomial expression at the k-th order. By using (2) and (3), the unknown coefficients can be estimated by solving the following MSE equation under the constraint for minimizing the difference between $y(s)$ in (2) and $\hat{y}(s)$ in (3).

$$\Gamma = \arg\min_{\beta_k} \left[\sum_{s} \left| y(s) - \sum_{k=0}^{L-1} \beta_k \cdot s^k \right|^2 \right] \tag{4}$$

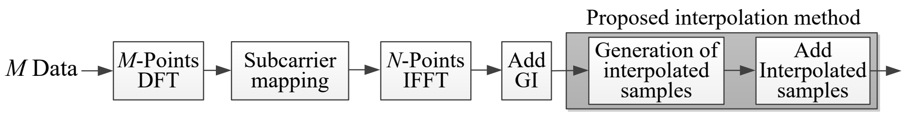
where s is taken by $2P+1$ samples as the reference samples which include $s=1$ to P at the last of $c(i\text{-}1, n)$, $s=P+Q+1$ to $2P+Q$ at the start of $c(i, n)$, and $s=m$ at the middle point. Here it should be noted that the middle point $y(m)$ is used to avoid the occurrence of higher peak amplitude within the interpolated Q samples which can keep a lower PAPR. The solution for β_k in (4) can be given by solving the following simultaneous equations.

$$[\beta_k]_{L\times 1} = \dagger[D(s,k)]_{(2P+1)\times L} \cdot [y(s)]_{(2P+1)\times 1} \tag{5}$$

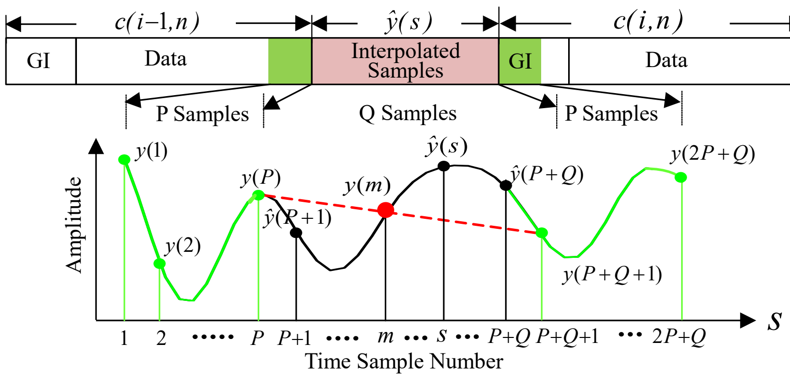
where $\dagger$ represents the Moore-Penrose inverse operation for $[D(s,k)]$ matrix with the size of $(2P+1)\times L$ which is given by,

$$D(s,k) = s^k, \ 1 \le s \le P, \ s = m, \ P+Q+1 \le s \le 2P+Q, \ and \ 0 \le k \le L-1 \tag{6}$$

From (6), it is obvious that the Moore-Penrose inverse operation for $[D(s,k)]$ can be calculated in advance because all elements of matrix in (6) are known at the transmitter which leads the considerable reduction of computation complexity in the estimation of unknown coefficients at every symbol. By using the estimated coefficients β_k in (5), the interpolated Q samples $\hat{y}(s)$ between two symbols from $s=P+1$ to $P+Q$ can be given by (3). Then Q interpolated samples are inserted between symbols.



(a) Structure of proposed DFTS-OFDM transmitter.



(b) Frame format and schematic diagram for proposed method.

Fig.1 Proposed spectrum regrowth reduction method for DFTS-OFDM.

At the receiver, the interpolated samples together with the GI inserted at the transmitter can be removed from the received time domain signal by using the detected symbol timing. The following processes for the equalization and demodulation are completely the same as that for the conventional DFTS-OFDM.

3 Performance evaluation for proposed method

This section presents simulation results for the proposed method as comparing with the conventional methods [2-4] which can realize the transceiver as almost the same processing load as that for DFTS-OFDM. In the simulations, the number of DFT points (M) and IFFT points (N) are 864 and 4096, the length of GI (N_g) is 256, allocated bandwidth (W) is 5MHz and the modulation method is 64QAM all of which parameters are the same as used in [2]. As for the non-linear amplifier, the SSPA with the Rapp's model is employed in which the smoothness factor to decide the non-linear level is 4 for the AM-AM conversion characteristics [6]. In the evaluation for the degree of spectrum regrowth reduction, this paper employs the spectrum efficiency SP_{EF} (bit/s/Hz) which is defined by the ratio of maximum transmission data rate V_{MAX} (bit/s) achieved in the allocated bandwidth (W) to the required channel spacing $F_{S\text{-}REQ}$ (Hz) in the operation of DFTS-OFDMA. $F_{S\text{-}REQ}$ (Hz) can be decided by using the adjacent channel leakage ratio (ACLR) at least 45dB [7]. V_{MAX} (bit/s) can be given by the multiplication of transmission efficiency and allocated bandwidth (W) which are given by NW/N_S for the conventional DFTS-OFDM, $(M-2)NW/MN_S$ for the phase-anchored, $(M-N_h-N_t)W/M$ for the zero-tail, $NW/(N_S+2E_X)$ for the filtering and $NW/(N_S+Q)$ for the proposed method. Here N_S is $(N+N_g)$, E_X is the length of extended GI samples for the filtering method and N_h and N_t are the numbers of zero-head and zero-tail for the zero-tail method.

As for the proposed method, L, P and Q are decided by computer simulations so as to obtain the maximum SP_{EF} (bit/s/Hz). From the results, the maximum SP_{EF} (bit/s/Hz) is obtained with keeping lower PAPR when L=18, P=10 and Q=20. As for the filtering method, E_X is taken by E_X=[Q/2] so as to be the same transmission efficiency as the proposed method. As for the zero-tail method, the number of zero-tail N_t is fixed by 54 so as to be the same length of GI ($N_g = 256$) after N-points IFFT with the over sampling ratio (N/M). As for the zero-head N_h, although the power spectrum regrowth can be improved as increasing the number of N_h, V_{MAX} (bit/s) would be decreased due to the fact that zero-head N_h is the pure overhead [3]. In this paper, the optimum number of N_h is selected which can achieve the maximum SP_{EF} (bit/s/Hz) by computer simulations. In the following evaluations, the optimum number of N_h selected at the given SSPA input-back off (IBO) is shown in the figures and table.

Fig. 2(a) shows the complementary cumulative distribution function (CCDF) of PAPR for the proposed and conventional methods. From the figure, it can be observed that the proposed method shows almost the same PAPR performance as the conventional DFTS-OFDM. This is the reason that the middle point of interpolated Q samples is used as the reference sample to avoid the occurrence of higher peak amplitude within the interpolated Q samples. Figs. 2 (b) and (c) show the normalized PSD in the linear and non-linear channels when the IBO of SSPA is

-7dB. From the figure, it can be observed that the proposed method in the linear channel can achieve much lower spectrum regrowth than the conventional methods. In the non-linear channel, the proposed method can achieve lower spectrum regrowth than the conventional methods especially at around just outside of allocated bandwidth which could improve the performance of ACLR. Fig. 2 (d) shows the performance of ACLR at the SSPA IBO of -7dB when changing the channel spacing F_S from 5 to 6MHz. From the figure, it can be observed that the proposed method can achieve smaller F_S at the required ACLR=45dB than the conventional methods. Table 1 shows the spectrum efficiency SP_{EF} (bit/s/Hz) for all methods when changing the SSPA IBO. From the table, it can be observed that the proposed method can achieve higher SP_{EF} than the conventional methods especially at higher IBO of SSPA.

Here it should be noted that the proposed method is required to generate the interpolated samples for every symbol at the transmitter although the conventional zero-tail and phase-anchored methods are required no additional processing load. In the proposed method, however the order of additional processing load for the multiplications required in (3) and (5) for obtaining Q interpolated samples is $O[LQ+(2P+1)L]$=738 which is small enough because the Moore-Penrose inverse operation in (5) can be calculated in advance. From this fact, the additional processing load required at the transmitter for the proposed method would cause little matter in the implementation of practical transmitter.

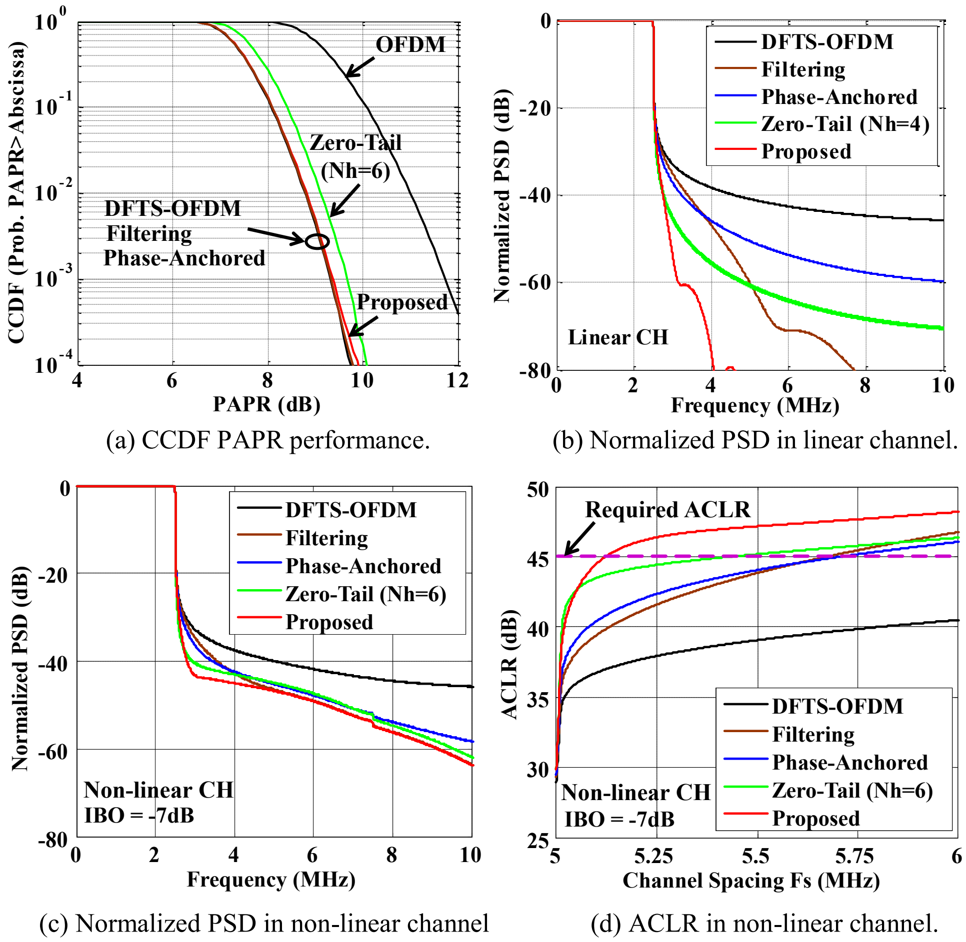


Fig. 2 Various performances for proposed and conventional methods.

Table 1 Comparisons of spectrum efficiency SP_{EF} (bit/s/Hz).

IBO of SSPA	DFTS-OFDM	Phase-Anchored	Filtering	Zero-Tail	Proposed
-6.0dB	NS	3.961	4.046	3.791 (N_h=2)	4.068
-7.0dB	NS	4.937	4.947	5.125 (N_h=6)	5.484
-8.0dB	NS	5.258	5.169	5.537 (N_h=4)	5.554
Linear	NS	5.297	5.222	5.559 (N_h=4)	5.565

NS=Not satisfy the required ACLR=45dB.

4 Conclusion

This paper proposed the simple power spectrum regrowth reduction method for DFTS-OFDM by inserting the interpolated samples between two time domain symbols. From the simulation results, this paper confirmed that the proposed method can achieve higher spectrum efficiency SP_{EF} (bit/s/Hz) in the non-linear channel than the conventional methods at the cost of slightly increasing of processing load at the transmitter.