OUT OF BAND AND ICI REDUCTION Methods

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Abstract:

As the demand for wireless communication systems increases, we need more frequency, bandwidth and the efficient and flexible use of existing bands. This problem can be solving by Cognitive Radio that use spectrum in a dynamic manner. A simple way for implementing this technology is OFDM (Orthogonal Frequency Division Modulation).but this modulation has some disadvantages like OOB (out of band) component, ICI (inter-carrier interference) and PAPR (peak to average power ratio), that in this paper we introduce some technique to reduce power of OOB and ICI.

Keyword – OFDM system, out of band (OOB) reduction, inter carrier interference (ICI).

I. INTRODUCTION

As the demand of wireless communication systems for high data rates rapidly increases, there is a substantial need for more frequency bandwidth and the efficient and flexible use of existing bands. Making use of the fact, that nowadays a significant part of the spectrum is unused due to the inflexible spectrum regulation, one approach to use the existing frequency gaps with a so-called overlay system is Cognitive Radio.

Orthogonal Frequency Division Modulation (OFDM) [1] has recently been applied widely in wireless communication systems due to its high data rate transmission capability with high bandwidth efficiency and its robustness to multi-path fading and delay channels. It has been used in digital video broadcasting (DVB) systems, wireless LAN standards such as American IEEE802.11a and the European equivalent HIPERLAN/2 and in multimedia wireless services such as Japanese Multimedia Mobile Access Communications. It has also been proposed as the core technique for the fourth generation (4G) mobile communications.

The main advantage of the OFDM system is its ability to convert a frequency selective fading channel into several nearly flat fading channels as the entire available spectrum is divided into a number of narrow band sub channels. The high spectral efficiency in the system is obtained by overlapping the orthogonal frequency responses of the sub channels.

One of the most challenging problems of any spectral sharing system is the successful coexistence with the licensed systems in the same frequency band, i.e., the spectral sharing system should not degrade the performance of the systems already working in that band. Considering OFDM based overlay systems this problem is reflected by significant out-of-band radiation caused by high side lobes of the modulated subcarriers. Several techniques for the reduction of the high out-of-band radiation already exist; that in this paper these techniques have been introduced.

Another main disadvantage of OFDM is its sensitivity against carrier frequency offset which causes attenuation and rotation of subcarriers, and intercarrier interference (ICI). The undesired ICI degrades the performance of the system. It is not possible to make reliable data decisions unless the ICI powers of OFDM systems are minimized. Thus, an accurate and efficient Intercarrier Interference (ICI) reduction procedure is necessary to demodulate the received data. Several methods have been presented that will introduce in this paper.

The paper is organized as follows. In Section II, the OFDM system model is described. In Section III the side lobe suppression techniques are explained. The ICI reduction methods are proposed in Section IV and the conclusion of this paper is given in Section V.

II. OFDM SIGNAL MODEL

In OFDM systems, the input high rate data stream is divided into many low-rate streams that are transmitted in parallel, thereby increasing the symbol duration and reducing the intersymbol interference (ISI).

An OFDM system with a total number of N subcarriers is considered. The input bits are symbol-mapped applying phase-shift keying (PSK) or quadrature amplitude modulation (QAM) and N complex-valued data symbols dn, $n = 1, 2, \ldots, N$, are generated and stacked into a data symbol array $\mathbf{d}=(d1, d2, \ldots, dN)T$, where (.)T denotes transposition. These symbols are serial to parallel (S/P) converted resulting in an N-element data symbol. Finally, the vector **d** is modulated on N subcarriers using the inverse discrete Fourier transform (IDFT). After that, parallel-to-serial (P/S) conversion is performed and a guard interval of length ΔT that exceeds the delay spread of the multipath channel is added as cyclic prefix.

Note that in this paper it is assumed that the guard interval is much shorter than the length $T0$ of the useful part of an OFDM symbol. Consequently, the influence of the guard interval on the spectrum of the OFDM transmission signal can be neglected and we assume, for simplicity reasons, $\Delta T \approx 0$ *in* the following.

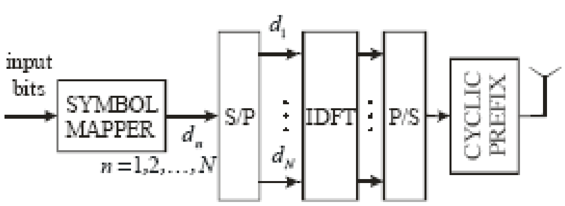

Fig. 1 OFDM system model

III. OUT OF BAND REDUCTION TECGNIQE

In the OFDM systems, N subcarriers are used for data transmission of N symbols dn. By using the IFFT to transform this OFDM signal from frequency to time domain a rectangular pulse shaping filter is implicitly applied. Thus, the spectrum of each individual subcarrier equals a sinc-function defined as $\mathrm{sinc}(x) = \sin(\pi x) / \pi x$ and is given as

$$Sn(x) = dn \cdot \mathrm{sinc}\,(x - xn),\ n = -N/2 \ldots N/2 - 1 \quad (1)$$

Where $x \in IR$ represents the frequency f shifted to the center frequency of the OFDM system f_0 and normalized to the sampling frequency $1/T0$. That $T0$ denoting the OFDM symbol duration excluding the guard interval and the normalized frequency is given as

$$x = (f - f_0)\, T_0 \quad (2)$$

Accordingly, $x_n = (f_n - f_0)\, T_0$ is defined as the normalized center frequency of the nth subcarrier with fn representing the center frequency of the nth subcarrier. The spectrum of the transmitted OFDM symbol is the superposition of the spectra of all individual subcarriers

$$s(x) = \sum_{-N/2}^{N/2-1} s_n(x). \quad (3)$$

The side lobe power of this sum signal only decays with $1/(x^2\ N)$ resulting in a high out-of-band radiation.

Here, some technique introduced to reduce the power of out of band component.

A1. Windowing

The interference between two sub channels that cause by out of band (OOB) components can be mitigated by windowing the OFDM signal in the time domain [2] [3].

In a simple OFDM system, symbols are performed using an N-FFT function. This implies that the received signal $r\,(k)$ is windowed in the time domain by a rectangular window function $w\,(k)$ resulting in:

$$\hat{r}\,(k) = r\,(k)\, w\,(k) \quad (4)$$

One possible countermeasure to overcome the interference is making the PDS of an OFDM modulated carrier (Sn(x)) go down more rapidly by windowing the transmit signal of the OFDM symbols. This makes the amplitude go smoothly to zero at the symbol boundaries. A commonly used window type is the raised cosine window that is defined by:

$$g(t)=\begin{cases}\frac{1}{2}+\frac{1}{2}\cos\left(\pi+\frac{\pi t}{\beta T_S}\right), & \text{for } 0\le t<\beta T_S\\ 1, & \text{for } \beta T_S\le t<T_S\\ \frac{1}{2}+\frac{1}{2}\cos\left(\frac{\pi(t-T_S)}{\beta T_s}\right), & \text{for } T_S\le t<(1+\beta)T_S\end{cases} \quad (5)$$

Where β denotes the rolloff factor and the symbol interval T_S is shorter than the total symbol duration $(1+\beta)\ T_S$ because adjacent symbols are allowed to partially overlap in the rolloff region.

Simulation shows that, unfortunately, the benefit of the raised cosine filter is very small for the first adjacent sub band, which the Interference power in first adjacent sub band as a function of β is almost constant curve. Even at very high rolloff factors, the achievable interference reduction is only about 6dB.

Analysis has shown that the benefit of raised cosine windowing technique with respect to interference reduction is fairly low and the drawback of this method is that windowing expands the signal in time domain and intersymbol interference is introduced.

A2. Guard band

One technique to reduce the side lobes is adaptive deactivation of adjacent subcarriers providing flexible guard bands between licensed and rental system [3] [4].

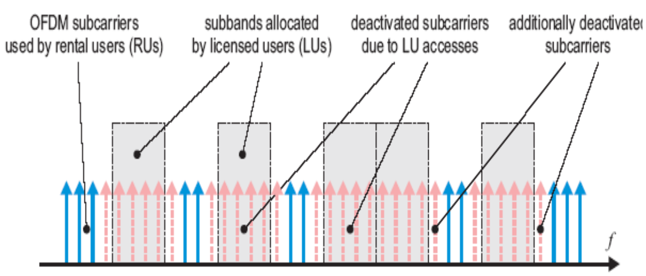

Fig. 2 Deactivation of adjacent subcarriers in guard band method

The drawback of this method is the less effective use of the available bandwidth.

Furthermore, a tradeoff between interference reduction and possible throughput has been presented.

Simulation results show that deactivation of the first adjacent subcarrier delivers the largest benefit and the additional deactivation of more subcarriers only provides a minor further improvement.

A3. Cancellation Carrier (CC)

OFDM systems exhibit significant out-of-band radiation caused by high side lobes of the modulated subcarriers. In this technique for reducing this undesirable effect, A few so-called cancellation carriers are inserted on the left and right hand side of the used OFDM spectrum[5][6]. These special subcarriers are not employed for data transmission, but carry complex weighting factors which are determined such that the side lobes of transmission signal and cancellation carriers cancel each other. (Note, similar approaches are used to reduce the peak-to-average-power ratio in OFDM systems.)

As depicted in Fig. 3 an OFDM system with N subcarriers is considered. The input bits are symbol-mapped onto N complex data symbols *dn* , n = −N/2, . . . , N/2 − 1. These symbols are serial-to-parallel (S/P) converted and fed into a side lobe suppression unit which inserts a few so-called CCs on the left and right hand side of the used OFDM spectrum. These special subcarriers carry complex weighting factors gm, m = 1, …, M, which are determined such that the side lobes of the CCs cancel the side lobes of the original Tx signal. The resulting symbol vector consisting of the complex data symbols *dn* and the weighting factors *gm* of the CCs is normalized such that the Tx power without CCs is equal to the signal power with CCs. After the insertion of the CCs the transmit symbol is modulated on N + M subcarriers using the inverse fast Fourier transform (IFFT) followed by a parallel-to-serial conversion (P/S).

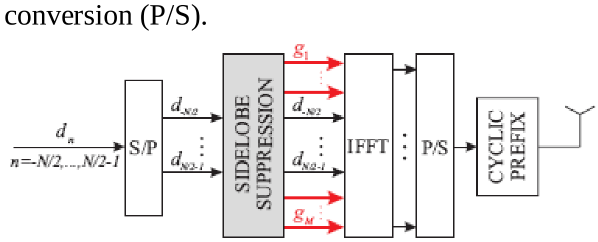

Fig. 3 Block diagram of the OFDM transmitter extended by the side lobe suppression unit for the insertion of CCs.

To suppress the high side lobes we insert *M* l and *M*r CCs on the left and on the right hand side of the used OFDM spectrum, respectively. Thus, the bandwidth of the OFDM spectrum is increased by $M = M_l + M_r$ Subcarriers to $N + M$ subcarriers. As the CCs are not used for data transmission the spectrum of the mth CC is not multiplied with a complex symbol and is weighted by 1 instead

$$c_m(x) = \mathrm{sinc}\,(\pi(x - y_m)),\ m = 1 \ldots M \tag{6}$$

The normalized center frequencies y_m of the M CCs lie on the left hand and the right hand side of the used spectrum, respectively and are given as

$$y_m = \begin{cases} (f_{-N/2-m} - f_0)T_0, & m = 1, \ldots, M_l \\ (f_{N/2-1+m-M_l} - f_0)\,T_0, & m = M_l + 1, \ldots, M \end{cases} \tag{7}$$

it is sufficient to reduce the spectrum to a certain number of sampling values. It is assumed, that K_l and K_r samples are considered in the optimization range on the left hand and on the right hand side, respectively. The total number of samples lying in the optimization range is defined as $K = K_l + K_r$. Thus, the spectrum of the Tx signal without CCs given in (3) is represented by K samples $s_k = s\,(w_k)$, $k = 1 \ldots K$, at the normalized frequencies w_k that are collected in the vector

$$\mathbf{s} = [s_1 \ldots s_K]^T \tag{8}$$

Accordingly,

$$\mathbf{c}_m = [c_{m,1} \ldots c_{m,K}]^T \tag{9}$$

Contains K samples of the spectrum of the mth CC from (6), i.e. $c_{m,k} = c_m(w_k)$

To perform the side lobe suppression each CC is multiplied by a *complex weighting factor* g_m, $m = 1 \ldots M$, which is determined such that the side lobes of the weighted CCs suppress the side lobes of the original Tx signal. That means that the superposition of the spectra of the weighted CCs and the original Tx signal has to be minimized. This optimization can be formulated as a linear least squares problem

$$\min_{\mathbf{g}} \left\| \mathbf{s} + \sum_{m=1}^{M} g_m \cdot \mathbf{c}_m \right\|^2 \quad \text{subject to} \quad \|\mathbf{g}\|^2 \le \alpha \tag{10}$$

With $g_m \in \mathbf{g} = [g_1 \ldots g_M]^T$. The constraint limits the power of the CCs to α and is added in order not to spend too much Tx power on the CCs. Finally, the resulting Tx signal in continuous frequency domain becomes

$$s'(x) = \sqrt{A} \cdot \left(s(x) + \sum_{m=1}^{M} g_m \cdot c_m(x) \right) \tag{11}$$

The factor $\sqrt{A}$ is introduced in order to keep the power of the signal without CCs the same as of the signal with CCs. Assuming that all complex symbols have a power normalized to $|d_n|^2 = 1$, can be rewritten and the normalization factor becomes

$$A\,(N + \|\mathbf{g}\|^2) = N \;\Leftrightarrow\; A = N / (N + \|\mathbf{g}\|^2) \le 1 . \tag{12}$$

The introduction of a normalization factor $A \le 1$ shows that a certain amount of the Tx power has to be spent on the CCs and is not available for data transmission leading to a loss in BER performance.

Simulation results show that with only two CCs at each side of the used spectrum the out-of-band radiation can be reduced by more than 20 dB. The price to pay for these results is an acceptable loss in BER performance and an increased computational complexity. But both effects can be kept at a minimum by appropriate countermeasures.

By inserting CCs the guard bands can be shortened and more subcarriers can be used for data transmission resulting in an increased spectral efficiency. The only drawbacks are a slight loss in bit error rate (BER) performance as a certain amount of the Tx power has to be spent on the CCs and the additional computational effort to determine the weighting factors.

A4. Subcarrier Weighting (SW)

This side lobe suppression technique is based on Subcarrier weighting [7] [8]. The real-valued subcarrier weights are determined in such a way that the side lobes of the transmission signal are minimized using an optimization algorithm which is capable to take several optimization constraints into account.

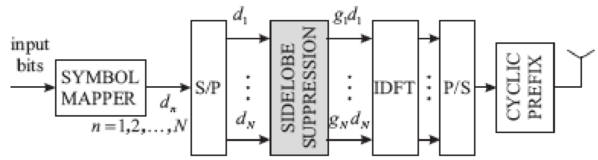

Fig. 4 Block diagram of the OFDM transmitter with subcarrier weighting.

An OFDM system with a total number of N subcarriers is considered. The block diagram of the OFDM transmitter is illustrated in Fig. 4. The input bits are symbol-mapped with phase-shift keying (PSK) modulation and N complex-valued data symbols d_n, with $|d_n|^2 = 1$, $n = 1, 2... N$, are generated. These symbols are serial-to-parallel (S/P) converted resulting in an N-element data symbol array $\mathbf{d} = (d_1, d_2... d_N)^T$. The array $\mathbf{d}$ is fed into the side lobe suppression unit which outputs are $\bar{\mathbf{d}} = (\bar{d}_1, \bar{d}_2 . . . \bar{d}_N)^T$. The side lobe suppression unit performs the multiplication of each symbol d_n with a *real valued weighting factor* g_n. Hence, the entries of $\bar{\mathbf{d}}$ are given by $\bar{d}_n = g_n * d_n$, $n=1, 2. . . N$. The weighting factors gn, are chosen such that the side lobes of the transmission signal are suppressed. Finally, the weighted vector $\bar{\mathbf{d}}$ is modulated on N subcarriers using the inverse discrete Fourier transform (IDFT). After that, parallel-to-serial (P/S) conversion is performed and a guard interval is added as cyclic prefix.

A single non-weighted subcarrier is $s_n(x)$, $n = 1, 2. . . N$, as our goal is to suppress the side lobes in certain frequency range, we consider $s_n(x)$ only in that range. We observe M normalized frequency samples y_m, $m = 1, 2. . . M$. which lie in the frequency range where the optimization of the side lobes is performed.

$$s_{n,m}=s_n(y_m)=d_n\frac{\sin\left(\pi(y_m-x_n)\right)}{\pi(y_m-x_n)}, \quad n=1,2,\ldots,N, \quad m=1,2,\ldots,M. \tag{13}$$

For simplicity, the optimization range is divided in two approximately equal parts which start from the first side lobes outside the OFDM transmission bandwidth.

Collecting $s_{n,m}$, into a vector we obtain $\mathbf{s}_n = (s_{n,1}, s_{n,2}, . . . , s_{n,M})^T$, $n = 1, 2, . . ., N$. Finally, stacking the vectors $\mathbf{s}_n$, into a matrix we get $\mathbf{S} = (\mathbf{s}_1, \mathbf{s}_2 . . . \mathbf{s}_N)$.

To minimize the side lobes of the weighted transmission signal $\bar{\mathbf{d}}$, we have to determine the vector $\mathbf{g}$ by solving the following optimization problem

$$\min_{\mathbf{g}} \|\mathbf{Sg}\|^2. \tag{14}$$

In addition, we include two constraints on the weighting vector $\mathbf{g}$.

1) Keeps the transmission power the same as in the case without weighting i.e. $\|\bar{\mathbf{d}}\|^2 = \|\mathbf{d}\|^2$.
2) Second constraint ensures that the elements of $\mathbf{g}$ are between pre-defined limits, i.e., $g_{min} \le g_n \le g_{max}$, $n = 1, 2. . . N$.

Such a constraint guarantees that each subcarrier receives a certain amount of the transmission power which is inherently controlled through the ratio $\rho = g_{max}/g_{min}$. Furthermore, g_{min} and g_{max} can be selected such that a weighted symbol $\bar{d}_n$ remains in the same decision region as the original symbol d_n as in such case no signaling from transmitter to receiver is required.

The optimization problem, together with those two constraints, can be solved numerically.

A possible drawback of the subcarrier weighting method is degradation in bit-error rate (BER) versus signal-to-noise ratio (SNR) performance as, due to the weighting, the subcarriers do not receive equal amounts of transmission power. The subcarrier weighting results in a BER loss, since the subcarriers do not receive equal amounts of the transmission power, it follows that if g_{max}/g_{min} grows, g_{min} becomes lower and thus, some subcarriers receive very small amounts of transmission power and cannot be decoded properly at the receiver resulting in performance degradation.

Numerical results show that side lobes can be easily suppressed by more than 10 dB with the proposed technique. But these results are achieved by allowing only a moderate loss in bit-error rate performance.

A5. Multiple Choice Sequences (MCS)

To reduce side lobes, we introduce another method referred to as multiple-choice sequences (MCS) [9]. This technique is based on the idea that transforming the original transmits sequence into a set of sequences. From this set, the sequence which offers the maximum reduction of out-of-band radiation is chosen for the actual transmission. To enable successful signal detection, de-mapping of the received sequence into the original sequence is required at the receiver. For this purpose, an index which uniquely identifies the selected sequence in the MCS set has to be signaled from transmitter to receiver. This results in a slightly reduced data throughput.

In this OFDM system model, the input bits are symbol-mapped applying phase-shift keying (PSK) or quadrature amplitude modulation (QAM) and N complex-valued data symbols dn, n = 1; 2; . . . ; N, are generated and stacked into a data symbol array d = (d_1; d_2; . . . ; d_N)^T . Then the array d is fed into the MCS side lobe suppression unit which outputs the sequence selected from the MCS set, denoted with ¯d = (¯d1; ¯d2; . . . ; ¯dN)^T, and the index of the chosen MCS, denoted with Q.

The principle of MCS is that, A set of P>1 sequences $d^{(p)}$ = ($d^{(p)}_1$; $d^{(p)}_2$; . . . ; $d^{(p)}_N$)^T, p = 1; 2; . . . ; P, is produced from the data sequence d. For each sequence, $d^{(p)}$ the average side lobe power, denoted with $A^{(p)}$, is calculated. To determine $A^{(p)}$, a certain frequency range spanning several OFDM side lobes, called optimization range, is considered using discrete frequency samples. Denoting the spectrum of an individual subcarrier with S(x), $A^{(p)}$ is given by

$$A^{(p)} = \frac{1}{K}\sum_{k=1}^{K}\left|\sum_{n=1}^{N} d_n^{(p)} S(y_k - x_n)\right|^2, p = 1, 2, \ldots, P \quad (15)$$

Where xn, n = 1; 2; . . . ; N, are the normalized subcarrier frequencies and yk, k = 1; 2; . . . ; K, are normalized frequency samples within the optimization range. The function S(x) depends on the applied transmit window and in the case of rectangular time domain windowing S(x) = sinc (x).

The index Q of the sequence with maximum side lobe suppression is given by

$$Q = \arg\min_{p} A^{(p)}, p = 1, 2, \ldots, P \quad (16)$$

Thus, the sequence $d^{(Q)}$ = ($d^{(Q)}_1$; $d^{(Q)}_2$; . . . ; $d^{(Q)}_N$)^T is output from the MCS unit, i.e. ¯d = $d^{(Q)}$

To enable successful data detection, the received sequence has to be de-mapped into the original sequence at the receiver. The MCS set is constructed such that the knowledge about the index Q of the selected sequence is sufficient to perform this de-mapping. Hence, the index Q is coded in bits, passed from the MCS unit to the signaling channel, and sent to the receiver.

Many practical MCS algorithms for generation of the MCS set can be derived. In the following, two simple algorithms to generate an MCS set are proposed.

- *Symbol constellation approach*

Symbol constellation algorithm generates the set of MCS such that the elements $d^{(p)}_n$, n = 1; 2; . . . ; N, of $d^{(p)}$ belong to the same symbol constellation as the elements of d. With this approach, the fact that different symbol sequences have side lobes with different powers is exploited.

Assume that the symbol constellation space consists of M points that are numbered as 0; 1; . . . ; M-1. To each data symbol dn, n = 1; 2; . . . ; N, an index $i_n \in${0; 1; . . . ; M-1} is assigned which corresponds to the number of the respective constellation point. Then, the index $i^{(p)}_n$ that corresponds to the MCS symbol $d^{(p)}_n$; n= 1; 2; . . . ; N, p = 1; 2; . . . ; P, is constructed by

$$i_n^{(p)} = \left((i_n + r_n^{(p)}) \bmod M\right) \quad (17)$$

That $r^{(p)}_n$ is an integer randomly chosen from the set $r^{(p)}_n \in$ {0; 1; . . . ; M-1}. After determining P index vectors $i^{(p)}$ = ($i^{(p)}_1$; $i^{(p)}_2$; . . . ; $i^{(p)}_N$)^T in the described manner, the MCS vectors $d^{(p)}$, p = 1; 2; . . . ; P, are obtained by taking the data symbols from the constellation space according to the vectors $i^{(p)}$. We assume that the same random generator and random seed for generating $r^{(p)}_n$, is used at both transmitter and receiver. Hence, the transformation of the received sequence back to the original sequence can be easily performed by exploiting the transmitted signaling information.

- *Phase approach*

In this approach, the MCS symbols are obtained by applying random phase shifts to the original symbols. Hence, the resulting MCS symbols are formed as

$$d_n^{(p)} = d_n \exp\left(j\varphi_n^{(p)}\right), \quad n = 1, 2, \ldots, N, \quad p = 1, 2, \ldots, P \quad (18)$$

Where the phase shifts $\varphi^{(p)}_n$ lie in the interval [0, 2π) and are generated as

$$\varphi_n^{(p)} = 2\pi\left(\frac{\bar{r}_n^{(p)}}{\bar{M}}\right) \quad (19)$$

¯M is a constant integer and ¯$r^{(p)}_n$ is an integer randomly chosen from the set ¯$r^{(p)}_n \in${0; 1; . . . M-1}. Thus, $\varphi^{(p)}_n$ can take one of the ¯M discrete phase values. Again, the same random seeds are used at the transmitter and receiver. Note that in the phase approach, the resulting MCS symbols do not necessarily belong to the same symbol constellation as the original symbols.

Numerical results show that with MCS approach OFDM side lobes can be reduced significantly while requiring only a small amount of signaling information to be sent from transmitter to receiver.

A6. Constellation Expansion (CE)

This algorithm is also use for reducing side lobe interference power levels in OFDM-based cognitive radios [10]. Exploiting the fact that different sequences have different side lobe power levels, this algorithm employs a constellation expansion based iterative approach in order to suppress the side lobe power levels. An important advantage of the proposed technique is that, no side information needs to be transmitted.

In the transmitter, the constellation expansion (CE) unit is added to a traditional OFDM system, which for a given OFDM input sequence determines the sequence that yields a low OOB radiation level. The inverse fast fourier transform (IFFT) is then applied to the new sequence.

In this technique, the symbols of a modulation scheme that modulates k bits/symbol and consisting of 2^k constellation points are mapped to a modulation scheme that modulates $(k+1)$ bits/symbol and consisting of 2^{k+1} constellation points. In other words, for every constellation point in the original symbol sequence, there are two points to choose from, in the expanded constellation space. Selecting one of the points on a random basis, each symbol in a sequence of N symbols is mapped to N symbols from the expanded symbol set. An underlying assumption with this technique is the transmitter and the receiver are assumed to have the knowledge of the points of the expanded constellation that are associated with the points in the original constellation. Hence, after the demodulation process, the symbols can be re-mapped to the points of the original constellation. With this knowledge, no side information is needed to be shared between the transmitter and the receiver.

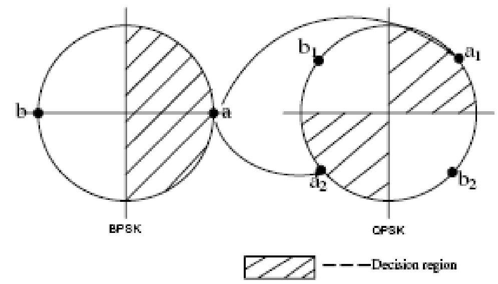

Fig. 5 a mapping approach for symbols from BPSK constellation to QPSK

As an example, an approach for mapping BPSK symbols to QPSK symbols is shown in Fig. 5. The rationale behind this association of points from a lower constellation to a higher constellation is to take advantage of the randomness involved in selecting one of the two points, and hence the combination of different in-phase and quadrature-phase components from all the subcarriers would result in a sequence with lower side lobes.

The proposed algorithm that selects the sequence randomly is shown in Fig. 6. For each symbol, the maximum interference power level is calculated and j iterations are performed. After each iteration, if the calculated interference power level of the new sequence is less than a pre-defined threshold or if the limit on the number of iterations is reached, the sequence with the lowest possible interference level out of all the randomly selected sequences is assigned to the original sequence. This process is repeated over all the symbols.

It can be noticed from the algorithm that the complexity of the algorithm is directly dependant on the value of *iterations threshold*. If the value of this variable is large, there is a greater probability of finding a sequence which has lower side lobe power levels.

The technique proposed in this section uses symbols from the higher order constellation diagram. Therefore, degradation in the BER and a slight degradation in the PAPR are expected.

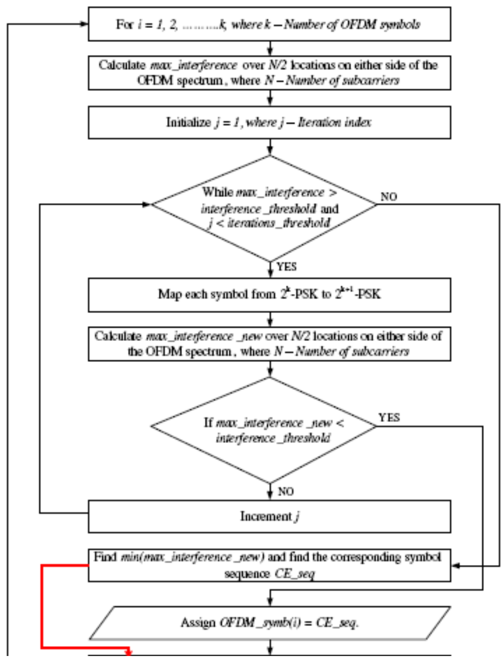

Fig. 6 The proposed algorithm for symbol selection using constellation expansion

A7. Additive Signal Method (AS)

This method, referred to as *additive signal* (AS) [11], is based on the addition of a complex valued sequence to the original transmit data sequence. This complex-valued sequence is determined according to an optimization algorithm that suppresses the side lobes and takes into account several constraints. In addition, this method can be easily combined with any of the existing side lobe suppression methods.

In this OFDM transmitter, the input bits are symbol-mapped applying phase-shift keying (PSK) or quadrature amplitude modulation (QAM) and N complex-valued data symbols d_n , $n = 1, 2, \ldots, N$, are generated. These symbols are serial to parallel (S/P) converted. Then array $\mathbf{d}$ is fed into the side lobe suppression unit which outputs $\mathbf{c} = (c_1, c_2 \ldots c_N)^T$. The side lobes suppression unit adds to each symbol d_n a complex-valued signal a_n. Hence, the entries of $\mathbf{c}$ are given by

$$c_n = d_n + a_n \quad n = 1, 2, \ldots, N \qquad (20)$$

The complex-values, a_n, $n = 1, 2\ldots N$, are chosen according to an optimization algorithm by which the side lobes of the transmission signal are suppressed. Finally, the vector $\mathbf{c}$ is modulated on N subcarriers using the inverse discrete Fourier transform (IDFT).

As our goal is to suppress the side lobes in certain frequency range, we consider $s_n(x)$ in equation (1), only in that range. We observe M normalized frequency samples, y_m, $m=1, 2\ldots M$ which lie in the frequency range where the optimization of the side lobes is performed. That

$$S_{n,m} = S_n(y_m) \quad n = 1, 2, \ldots, N, \quad m=1, 2, \ldots, M$$

Collecting, $S_{n,m}$, into a vector we obtain $s_n = (s_{n,1}, s_{n,2}, \ldots, s_{n,M})^T$. Finally, stacking the vectors s_n , into a matrix we get $S = (s_1, s_2, \ldots, s_N)$.

To keep the dimension of the matrix $\mathbf{S}$ low only one normalized frequency sample per side lobe is considered in the optimization range. In addition, these frequencies y_m are chosen such as to satisfy

$$\sin(\pi (y_m - x_n)) = 1, \; n = 1, 2\ldots N, \; m=1, 2\ldots M \qquad (21)$$

To minimize the side lobes of the transmission signal $\mathbf{c} = \mathbf{d} + \mathbf{a}$, we have to determine the vector $\mathbf{a}$ by solving the following optimization problem

$$\mathbf{a} = \min_{\tilde{\mathbf{a}}} \left\| \mathbf{S}(\mathbf{d} + \tilde{\mathbf{a}}) \right\|^2 \qquad (22)$$

Where $\tilde{\mathbf{a}}$ is a trial value of $\mathbf{a}$. Note that $\|S(d+\tilde{a})\|^2$ is the side lobe power of the transmission signal in the optimization range.

Also, we include two constraints on the vector $\mathbf{a}$. The first ensures that the sequence $\mathbf{c}$ does not invest more power into transmission than the original sequence $\mathbf{d}$, i.e., $\|d+a\|^2 \leq \|d\|^2$

The second constraints inherently controls bit-error-rate (BER) performance and ensure that the elements of $\mathbf{a}$ are between pre-defined limits, i.e., $\|a_n\| \leq R$, $n = 1, 2\ldots N$.

Numerical results show that there are two counteracting effects caused by the radius R. Enlarging this radius improves side lobe suppression, but simultaneously leads to a further loss in SNR performance. Therefore, there is a trade-off between the additional side lobe suppression obtained by enlarging the radius R and the increased loss in SNR performance.

This side lobe suppression scheme can be easily combined with existing side lobe suppression techniques such as transmit windowing and enables remarkable side lobes reduction by more than 10 dB. But the price to pay for this achievement is a moderate loss in BER performance.

A8. Combined Methods

Some techniques have been described above to reduce the power of out of band component. We can combine these methods to suppress the side lobes of OFDM transmission signal.

Some of them are:

- CC + CE
- CC + Windowing
- MCS + SW
- MCS + CC
- SW + Guard band

With these techniques the spectral efficiency of OFDM based transmission systems can be improved and this approach can be applied to OFDM based overlay system to avoid interference towards the legacy system sharing the same frequency band.

IV. ICI Reduction Methods

One of the main disadvantages of OFDM is its sensitivity against carrier frequency offset which causes attenuation and rotation of subcarriers, and intercarrier interference (ICI). This frequency offset causes by carrier frequency mismatch between the transmitter and receiver, and/or the Doppler shift.

The undesired ICI degrades the performance of the system. It is not possible to make reliable data decisions unless the ICI powers of OFDM systems are minimized. Thus, an accurate and efficient Intercarrier Interference (ICI) reduction procedure is necessary to demodulate the received data.

The signal at the output of the OFDM transmitter resulting from the i-th transmitted symbol is given by

$$x(t) = \exp(j2\pi f_c t) \sum_{k=0}^{N-1} b_{k,i} p\left(t - \frac{kT}{N}\right) \tag{23}$$

Where fc is the carrier frequency and p(t) is the impulse response of the low-pass filter in the transmitter. At the receiver, the signal is mixed with a local oscillator signal which is Δf above the correct frequency fc. Ignoring the effects of noise, the demodulated signal is then given by

$$y(t) = \exp j(2\pi \Delta f t + \theta_0) \sum_{k=0}^{N-1} b_{k,i} q\left(t - \frac{kT}{N}\right) \tag{24}$$

Where q(t) is the combined impulse response of the channel and of the transmitter and receiver filters.θ_0 is the phase offset between the phase of the receiver local oscillator and the carrier phase at the start of the received symbol.

After sampling and DFT, the received signal on subcarrier k can be written as

$$Y(k) = X(k)S(0) + \sum_{l=0,\, l\neq k}^{N-1} X(l)S(l-k) + n_k, \quad k = 0,\, 1,\, \ldots,\, N-1 \tag{25}$$

Where N is the total number of the subcarriers denotes the transmitted symbol for the k-th subcarrier and n_k is an additive noise sample. The sequence *S (l-k)* is defined as the ICI coefficient between *l*-th and *k*-th subcarriers, which can be expressed as

$$S(l-k) = \frac{\sin(\pi(l+\varepsilon-k))}{N \sin\left(\frac{\pi}{N}(l+\varepsilon-k)\right)} \cdot \exp\left(j\pi\left(1-\frac{1}{N}\right)(l+\varepsilon-k)\right) \tag{26}$$

Where $\varepsilon = f_D / \Delta f$ is the normalized frequency offset, f_D is Doppler frequency and Δf is subcarrier frequency separation.

The first term in the right-hand side of (25) represents the desired signal. Without frequency error (ε =0), S (0) takes its maximum value. The second term is the ICI components, which as ε becomes larger, the desired part |S (0)| decreases and the undesired part |*S(l-k)*| increases.

The system ICI power level can be evaluated by using the CIR (Carrier-to-Interference power Ratio). While deriving the theoretical CIR expression, the additive noise is omitted. It is assumed that the transmitted data have zero mean and are statistically independent; therefore, the CIR expression for subcarrier can be derived as

$$\mathrm{CIR} = \frac{|S(k)|^2}{\sum_{l=0,\, l\neq k}^{N-1} |S(l-k)|^2} = \frac{|S(0)|^2}{\sum_{l=1}^{N-1} |S(l)|^2}. \tag{27}$$

Therefore, the CIR of OFDM systems only depends on the normalized frequency offset ε approximately.

Here, some methods introduced to reduce the power of intercarrier interference (ICI).

B1. ICI self-cancellation

Equation (10), (11) shows that both real and imaginary parts of the ICI coefficient are gradually changed with respect to the subcarrier index. For the majority of *l-k* values, the difference between *S (l-k)* and *S (l-k+1)* is very small. Therefore, if a data pair (a,-a) is modulated onto two adjacent subcarriers *(l, l+1)*, where *a* is a complex data, then the ICI signals generated by the subcarrier will be cancelled out significantly by the ICI generated by subcarrier *l+1*. This is the ICI cancellation idea in this method. [12] [13]

Assume the transmitted symbols are constrained so that X (1) = - X (0), X (3) = - X (2)… X (N-1) = - X (N-2), then the received signal on subcarrier k becomes

$$Y'(k) = \sum_{\substack{l=0 \\ l=\text{even}}}^{N-2} X(l)\left[S(l-k) - S(l+1-k)\right] + n_k \tag{28}$$

In such a case, the ICI coefficient is denoted as

$$S'(l-k) = S(l-k) - S(l+1-k) \tag{29}$$

For most of the *l-k* values, it is found that |S´ (l-k)| << |S (l-k)|.

The signal redundancy makes it possible to improve the system performance at the receiver side, which we can compute as

$$\begin{aligned} Y''(k) &= Y'(k) - Y'(k+1) \\ &= \sum_{\substack{l=0 \\ l=\text{even}}}^{N-2} X(l)[-S(l-k-1) + 2S(l-k) \\ &\qquad - S(l-k+1)] + n_k - n_{k+1} \end{aligned} \tag{30}$$

The corresponding ICI coefficient then becomes

$$S''(l-k) = -S(l-k-1) + 2S(l-k) - S(l-k+1) \tag{31}$$

For the majority of *l-k* values, |S´ (l-k)| is much smaller than |S (l-k)|, and the |S´´ (l-k)| is even smaller than |S´(l-k)|. Thus, the ICI signals become smaller when applying ICI cancelling modulation. On the other hand, the ICI cancelling demodulation can further reduce the residual ICI in the received signals. This combined ICI cancelling modulation and demodulation method is called the ICI self-cancellation scheme.

Due to the repetition coding, the bandwidth efficiency of the ICI self-cancellation scheme is reduced by half. To fulfill the demanded bandwidth efficiency, it is natural to use a larger signal alphabet size. For example, using 4PSK modulation together with the ICI self-cancellation scheme can provide the same bandwidth efficiency as standard OFDM systems (1 bit/Hz/s).

B2. Pulse shaping

The N-subcarrier OFDM block with pulse-shaping is expressed as

$$x(t) = e^{j2\pi f_c t} \sum_{k=0}^{N-1} a_k p(t) e^{j2\pi f_k t} \tag{32}$$

Where f_c is the carrier frequency, *fk* is the subcarrier frequency of the *k*th subcarrier, p (t) is the time-limited pulse shaping function, and a_k where k= 0, 1… N-1 is the data symbol transmitted on the kth subcarrier. We assume that a_k has mean zero and normalized average symbol energy. [14] [15]

A number of pulse shaping functions such as Rectangular pulse (REC), Raised cosine pulse (RC), Better then raised cosine pulse (BTRC), Sinc power pulse (SP) and Improved sinc power pulse (ISP) have been introduced for ICI power reduction. The functions are defined as below:

- Rectangular pulse

$$P_{REC}(f) = \sin c(fT) \tag{33}$$

- Raised cosine pulse

$$P_{RC}(f) = \sin c(fT)\frac{\cos(\pi\alpha fT)}{1-(2\alpha fT)^2}, \tag{34}$$

- Better then raised cosine pulse

$$P_{BTRC}(f) = \sin c(fT)\frac{2\beta fT\sin(\pi\alpha fT) + 2\cos(\pi\alpha fT) - 1}{1+(\beta fT)^2} \tag{35}$$

- Sinc power pulse

$$P_{SP}(f) = \sin c^n(fT) \tag{36}$$

- Improved sinc power pulse

$$P_{ISP}(f) = \exp\{-a(fT)^2\}\operatorname{sinc}^n(fT). \tag{37}$$

Where α(0≤α≤1) is the roll off factor and $\beta = \pi\alpha/\ln 2$, '*a'* is a design parameter to adjust the amplitude and *n* is the degree of the sinc function.

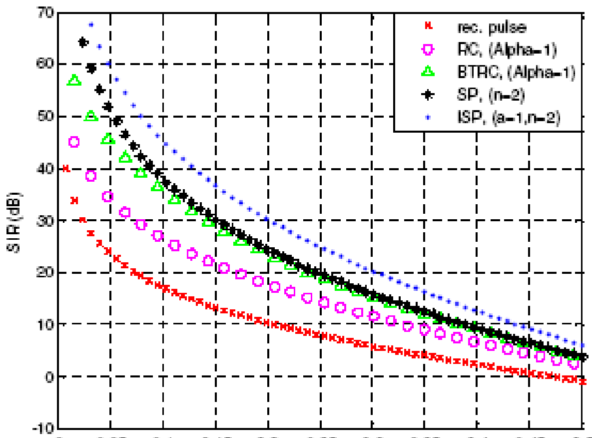

Fig. 7 CIR performance for different pulse shapes

The purpose of pulse shaping is to reduce the side lobe, as the side lobe contains the ICI power. Fig. 7 shows that the side lobe is maximum for rectangular pulse and minimum for ISP pulse shapes. This property of ISP pulse shape will provide better performance in terms of ICI reduction than those of the other pulse shapes.

B3. Adaptive Select Mapping

This technique is based on partitioning a (2M)-point constellation into two disjoint constellations and adaptively mapping each $\log_2$ M input bits into either of these two constellations. The consequent extra freedom can be exploited to reduce the PICR. A great advantage of this technique is that no side information is required at the receiver. [16]

If the carrier frequency offset between the transmitter and receiver exists, the receiver output becomes (25).

It is evident that for an arbitrary signal constellation, some data sequences lead to low CIR. Based on this fact; one can reduce the CIR by mapping those data sequences having high CIR to some data sequences having low CIR.

Having above considerations, one can map an M-point signal constellation *QM* to a (2M)-point signal constellation *Q2M*. Then, each point in *QM* can be represented by either of the two corresponding points in Q2M. That is, each OFDM subcarrier has two modulation choices. The consequent extra freedom allows carefully selecting each subcarrier modulation symbol to reduce the CIR leading to the adaptive mapping technique.

Various schemes can be used to map *QM* to *Q2M*. Here, the following scheme that only involves sign-changing is used. Given an origin-symmetrical signal constellation *Q2M* (e.g. 2MPSK or 2M-QAM), one can partition it into two disjoint M-point constellations such that $Q2M = Q^{(1)}M \cup Q^{(2)}M$

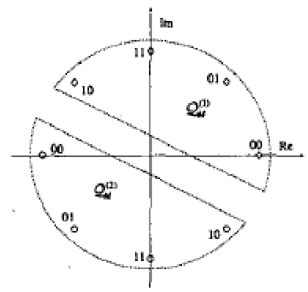

Fig. 8 Partitioning BPSK into two disjoint constellations

There are some methods to find suboptimal mapping sequences with significantly lowered computational complexity

- Random selection
- Modified partial Transmit Sequence (PTS)
- Recursive partial sequence (RPS)

There are some other methods to reduce the ICI power like Time-domain windowing [17], Frequency-domain equalization [18], FrFT based OFDM. That reader is referred to the papers listed at the References.

V. Conclusion

In this paper, we introduce some methods for reducing the power of OOB and ICI as most challenging problems of OFDM.

By comparing these techniques for OOB and ICI (and PAPR) power reduction, we understand that some of them are similar. For example Multiple Choice Sequence is use for all of them. So, maybe we can use it by an optimization problem and some constraint to reduce the power of OOB, ICI and PAPR together.

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