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**Title:** Generalized Precoded OFDMA (GPO) – a SCFDMA variant for achieving near unit PAPR and low out of band emission  
**Agenda Item:** 7.2.6  
**Document for:** Discussion

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## 1. Introduction

In RP-151261, it is agreed that the following will be supported for NB-IOT operation:

- 180 kHz UE RF bandwidth for both downlink and uplink
- OFDMA on the downlink
  - Two numerology options will be considered for inclusion: 15 kHz sub-carrier spacing (with normal or extended CP) and 3.75 kHz sub-carrier spacing. Technical analysis will either perform a down-selection or decide on inclusion of both based on the feasibility of meeting relevant requirements while achieving commonality (to be finalized by RAN #70)
- For the uplink, two options will be considered: FDMA with GMSK modulation (as described in 3GPP TR 45.820 section 7.3), and SC-FDMA (including single-tone transmission as a special case of SC-FDMA)

In this contribution, we introduce a variant of SC-FDMA modulation that is proposed in [7]. The current proposal has the following characteristics:

- a) The waveform can be used with either 15 KHz subcarrier spacing 3.75 KHz subcarrier spacing
- b) The waveform achieves very low PAPR for both low and large subcarrier allocations.
- c) The waveform proposed in [7] gives low PAPR using  $\pi/2$  BPSK. The modified frequency domain filter introduced in this proposal further reduces the PAPR. We obtain a PAPR close to unity when certain excess band width is allowed. Very low PAPR is achieved even without using excess BW.
- d) Using a raised cosine window, the proposed waveform offers very low out of band emissions (OBE)
- e) The waveform requires a frequency domain widely linear receiver whose computational complexity is similar to conventional SC-FDMA receivers

## 2. Waveform Design

A Generalized- precoded-OFDM (GPO) waveform with a very low peak-to-average-power-ratio (PAPR) is proposed. In this method, data of multiple users are multiplexed in frequency domain using orthogonal or non-orthogonal subcarrier mapping. The method comprises of: constellation rotation of user data, DFT precoding and spreading, user specific frequency domain pulse shaping (FDPS) possibly with certain excess bandwidth. The frequency domain pulse shaping filter (FDPSF) introduces inter-symbol-interference (ISI) and possibly inter-user interference (IUI). For the special case of binary modulation, a waveform with low PAPR is obtained using frequency domain pulse shaping filters (FDPSFs) that are derived from the linearized Gaussian pulse. To mitigate the ISI and self-interference caused by the FDPSF, widely linear frequency domain equalization method is used that has low-implementation complexity.

### Waveform Description

Consider the waveform design for a single user case. Generalization to the case of multiple users simultaneously sharing the available bandwidth will follow. The transmitter sends a block of  $M$  i.i.d real/complex valued modulation

alphabets with zero-mean and variance  $\sigma^2$ . Let  $a_t(l)$  denote the modulation data. We apply a constellation specific phase rotation  $\theta(l)$  to obtain:  $x_t(l) = e^{j\theta(l)}a_t(l)$ . The DFT precoding of the data stream  $x_t(l)$  is accomplished using a  $M$ -point DFT as

$$x(k) = \sum_{l=0}^{M-1} x_t(l) e^{-j2\pi lk/M} \quad k = 0, \dots, M-1 \quad (1)$$

where  $l, k$  denote the discrete time and subcarrier indices, respectively, and  $x(M+k) = x(k)$ . Alternative to (1), a two sided DFT can be taken as.

$$x(k) = \sum_{l=-\frac{M}{2}}^{\frac{M}{2}-1} e^{j\theta_t(l)} x_{t,l}(l) e^{-j2\pi lk/M} \quad -\frac{M}{2} \leq k \leq \frac{M}{2}-1$$

Consider a  $L$  fold periodic extension of  $x(k)$  where  $\tilde{x}(m) = x\left(\left(\left(m + \frac{LM}{2}\right) \bmod M\right) + 1\right)$ . Here, the elements of the vector  $\tilde{x}(m)$  take the range  $m = -\frac{LM}{2}, \dots, \frac{LM}{2}-1$  and  $ML = N, N$  being total number of used

subcarriers. In time domain,  $\tilde{x}_t(n) = x_t\left(\frac{n}{L}\right)$  for  $n = pL$  where  $p = 0, 1, \dots, M-1$  and  $\tilde{x}_t(n) = 0$  elsewhere and  $n = -\frac{N}{2}, \dots, \frac{N}{2}-1$ . Here  $x_t(l) = \tilde{x}_t\left(lL - \frac{N}{2}\right)$  for  $l = 0, \dots, M-1$ . Let

$$\tilde{x}(m) = \sum_{n=-\frac{N}{2}}^{\frac{N}{2}} \tilde{x}_t(n) e^{-\frac{2\pi mn}{N}} \quad m = -\frac{N}{2}, \dots, \frac{N}{2}-1 \quad (2)$$

$$= \sum_{l=0}^{M-1} x_t(l) e^{-\frac{j2\pi(lL - \frac{N}{2})m}{N}} \quad (3)$$

$$= e^{j\pi m} \sum_{l=0}^{M-1} x_t(l) e^{-\frac{j2\pi lLm}{N}}$$

(4)

On line 1 we substitute the variable  $n$  with  $(lL - \frac{N}{2})$ .

**Note:** Note that the DFT operation in (1) can also be implemented as a two sided DFT with  $l$  in the range  $\left[-\frac{M}{2}, \dots, \frac{M}{2}-1\right]$ . Alternatively, swap the left and right halves of  $x(k)$  with zero frequency component in the middle.

Now consider a frequency domain pulse shaping filter

$$q(m) = \sum_{n=-\frac{N}{2}}^{\frac{N}{2}-1} q_t(n) e^{-\frac{j2\pi nm}{N}}, \quad m = -\frac{N}{2}, \dots, \frac{N}{2}-1 \quad (5)$$

Where  $q_t(n)$  are the samples of the time domain pulse shaping filter. Note that  $q(m)$  may take zero values for certain subcarriers. Alternatively, all  $N$  subcarriers need not be modulated with data. In some cases, some subcarriers at band edges may be nulled out. Applying the pulse shape to the transmitted data  $\tilde{x}(m)$ , we have:  $x_q(m) = q(m)\tilde{x}(m)$ . The time domain baseband signal  $s(t)$  is obtained using an inverse discrete time Fourier transform (IDFT).

$$s(t) = \frac{1}{N} \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} q(m) \tilde{x}(m) e^{j2\pi m \Delta f (t-T_{CP})}, \quad t \in [0, T + bT] \quad (6)$$

where  $T$  is the useful portion of OFDMA symbol,  $T_{CP}$  is the duration of the cyclic prefix (CP) and  $\Delta f = \frac{1}{T}$  is the subcarrier spacing. Note that  $b=1$  when the system uses CP only and  $b=2$  when the system uses cyclic prefix as well as cyclic suffix. Using (4) and (5), the analog signal can be rewritten as

$$s(t) = \frac{1}{N} \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} e^{j\pi m} q(m) \tilde{x}(m) e^{j2\pi m \Delta f (t-T_{CP})}, \quad t \in [0, T + bT] \quad (7)$$

$$= \frac{1}{N} \sum_{l=0}^{M-1} x_t(l) \sum_{m=-\frac{LM}{2}}^{\frac{LM}{2}-1} q(m) e^{j2\pi m (\frac{1}{T}(t-T_{CP}) - \frac{lL}{N} + \frac{1}{2})} \quad (8)$$

$$= \frac{1}{N} \sum_{l=0}^{M-1} e^{j\theta(l)} a_t(l) q_p(t - T_{CP} - \frac{lT}{M} + \frac{T}{2}) \quad (9)$$

$$q_p(t) = \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} q(m) e^{j2\pi m \frac{t}{T}}$$

where  $x_t(l) = e^{j\theta(l)} a_t(l)$ . Note that  $q_p(t) = q_p(t + rT)$ ,  $r$  being an integer. The transmitter sends successive data blocks serially where each data block is limited to a duration of  $T + bT_{CP}$  seconds. Here, the time domain signal has a form similar to conventional SC-FDMA with  $q(t)$  being the pulse shaping function and  $t \in [0, T + bT]$  is the time domain pulse shaping function and

### Multiple Access

Consider the multiple access scenarios where a number of users share the available bandwidth simultaneously. We consider both non-orthogonal and orthogonal user allocations where the users may employ distinct pulse shapes with different bandwidth requirements. Assuming there are a total of  $u$  users, let us denote the data of the  $i^{\text{th}}$  user with  $x_i(k)$  where

$$x_i(k) = \sum_{l=0}^{M_i-1} e^{j\theta_i(l)} x_{i,t}(l) e^{-\frac{j2\pi lk}{M_i}}, \quad k = 0, \dots, M_i - 1 \quad (10)$$

where  $M_i$  is the data length of the  $i^{\text{th}}$  user,  $\theta_i(l)$  being the constellation rotation employed by the  $i^{\text{th}}$  user data  $x_{i,t}$ . Note that the DFT operation in (10) can be implemented using a two sided DFT as

$$x_i(k) = \sum_{l=-\frac{M_i}{2}}^{\frac{M_i}{2}-1} e^{j\theta_i(l)} x_{i,t}(l) e^{-\frac{j2\pi lk}{M_i}} \quad -\frac{M_i}{2} \leq k \leq \frac{M_i}{2} - 1$$

Let  $\tilde{x}_i(m)$  denote the  $L_i$  fold periodic extension of  $x_i(k)$  where  $L_i M_i = N$  and let  $q_i(l)$  be the FDPSF associated with this user that is defined as

$$q_i(m) = \sum_{n=-\frac{N}{2}}^{\frac{N}{2}-1} q_{i,t}(n) e^{-\frac{j2\pi mn}{N}}, \quad \text{for } m = -\frac{\bar{M}_i}{2}, \dots, \frac{\bar{M}_i}{2} - 1 \quad (11)$$

$$= 0 \quad \text{elsewhere} \quad (12)$$

where  $q_{i,n}(t)$  are the corresponding time domain samples. Note that the FDPSF takes non-zero values over  $M_i$  subcarriers where  $M_i - M_i$  is the excess number of subcarriers employed for the  $i^{\text{th}}$  user. In this case,  $(M_i - M_i)\Delta f$  is denoted as the excess bandwidth employed by the  $i^{\text{th}}$  user. Further, the users are frequency multiplexed over the given band of interest as

$$s_i(t) = \frac{1}{N} \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} q_i(m - m_i) \tilde{x}_i(m) e^{j2\pi\Delta f(t - T_{CP})}, \quad t \in [0, T + bT_{CP}] \quad (13)$$

where  $m_i$  is the frequency shift of the  $i^{\text{th}}$  user. The proposed method results in a non-orthogonal multicarrier signal if the values of  $m_i$  are set to integer multiples of  $\bar{M}_i$ . Note that, the values of  $m_i$  are chosen based on the subcarrier mapping procedure employed by the system.

The transmitted signal can be represented in an alternative form as

$$s_i(t) = \frac{1}{N} \sum_{l=0}^{M_i-1} e^{j\theta_i(l)} a_{i,t}(l) q_{i,p}(t - T_{CP} - \frac{lT}{M_i} + \frac{T}{2}), \quad t \in [0, T + bT_{CP}] \quad (14)$$

$$q_{i,p}(t) = \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} q_i(m - m_i) e^{j2\pi m \frac{t}{T}}$$

Where  $q_{i,p}(t)$  is the time domain pulse shaping function employed by the  $i^{\text{th}}$  user.

Here,  $q_{i,p}(t) = q_{i,p}(t + rT)$ , where  $r$  is an integer.

Let

$$q_{i,p}(t) = \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} q_i(m - m_i) e^{j2\pi m \frac{t}{T}} \quad t \in [0, T + bT_{CP}] \quad (15)$$

Using (11) and substituting  $m - m_i = m_1$  in (15) we express  $q_i(t)$  in alternative form as

$$q_{i,p}(t) = e^{j2\pi m_i \frac{t}{T}} \sum_{m=-\frac{\bar{M}_i}{2}}^{\frac{\bar{M}_i}{2}-1} q_i(m) e^{j2\pi m \frac{t}{T}} \quad (16)$$

$$= q_{0,i}(t) e^{-j2\pi m_i \frac{t}{T}} \quad t \in [0, T + bT_{CP}] \quad (17)$$

$$q_{0,i}(t) = \sum_{m_1=-\frac{\bar{M}_i}{2}}^{\frac{\bar{M}_i}{2}-1} q_i(m_1) e^{j2\pi m_1 \frac{t}{T}}$$

and

$q_{0,i}(t)$  is the baseband pulse shaping function used by the  $i^{\text{th}}$  user. Now we can rewrite the transmitted signal as

$$s_i(t) = \frac{1}{N} \sum_{l=0}^{M_i-1} e^{j\theta_i(l)} a_{i,t}(l) q_{0,i}(t - T_{CP} - \frac{lT}{M_i} + \frac{T}{2}) e^{j2\pi m_i \frac{(t - T_{CP} - \frac{lT}{M_i} + \frac{T}{2})}{T}} \quad (18)$$

In practice, the total number of subcarriers is  $N$ . However, only  $N_u$  subcarriers out of  $N$  may be used by the system. The remaining  $(N-N_u)$  subcarriers do not carry the data. Furthermore, a number of users are frequency multiplexed over the  $N_u$  subcarriers. The transmitted signal in its general form is given by

$$s_i(t) = \frac{1}{N_u} \sum_m q_i(m - m_i) \tilde{x}_i(m) e^{j2\pi\Delta f(t - T_{cp})} \quad t \in [0, T + bT_{cp}]$$

Here the transmitted signal spans over a group of subcarriers whose range is dictated by the subcarriers occupies by the signal of the  $i$ th user. Furthermore, the value of  $m_i$  is a system design feature that can be used to control the

amount of non-orthogonality introduced by the system. The value of  $m_i$  may be set to  $M_i$ ,  $\bar{M}_i$  or any other value.

For example, setting the value of  $m_i$  in the range  $[0, M_i]$  increases the spectrum efficiency of the system. Another

alternative is to choose the value of  $m_i$  in the range  $[M_i, \bar{M}_i]$ . Note that the modulation type, the DFT size, and the type of FDPSF used can be varied at user level.

### Symbol Windowing

As the transmit signal is confined to a period of one OFDM symbol duration, effectively it imposes a rectangular window function that leads to high OBE. In order to reduce OBE we employ alternative time domain window functions that offer smooth transitions at the OFDM symbol boundaries. The proposed method includes a cyclic prefix as well as cyclic postfix each of duration  $T_{CP}$ . The analog signal can be rewritten as

$$s_i(t) = \frac{1}{N} \sum_{m=-\frac{N}{2}}^{\frac{N}{2}-1} w(t) q_i(m - m_i) \tilde{x}_i(m) e^{j2\pi m \Delta f (T - T_{CP})}, \quad t \in [0, T + bT_{CP}] \quad (19)$$

$$= \frac{1}{N} \sum_{l=0}^{M_i-1} e^{j\theta_i(l)} a_{i,t}(l) w(t) q_{i,p}(t - T_{CP} - \frac{lT}{M_i} + \frac{T}{2}), \quad t \in [0, T + bT_{CP}] \quad (20)$$

where  $w(t)$  is the proposed window function defined over the interval  $t \in [0, T + bT_{CP}]$  (that is designed as the OFDM symbol block duration). It is common practice to choose the window  $w(t)$  such that it takes a constant value for the duration of the OFDM symbol that excludes cyclic prefix and suffix. Additionally, the window may take a constant value during a portion of the cyclic prefix and/or suffix and it tapers to a zero value at the block boundaries. Raised cosine (RC) window is used in this contribution.

**Note:** To avoid the dc subcarrier, a half subcarrier shift can be introduced as in the case of SC-FDMA.

### Frequency Domain Pulse Shaping Filter

This section proposes a design procedure that enables low PAPR waveform design. The PAPR can be controlled using a suitable choice of constellation rotation factor  $\theta(l)$ , modulation size  $Q$  and the FDPSF  $q(m)$  given in (5). We

let  $\theta(l) = \frac{\pi(l-1)}{2}$  for real constellations, and for  $Q$ -ary complex constellations such as QAM, we let

$\theta(l) = \frac{\pi(l-1)}{Q}$ . For the special case of  $Q=2$  (that is the focus of this SI), we design waveforms with nearly constant

envelope by selecting  $\theta(l) = \frac{\pi(l-1)}{2}$ , and by choosing the FDPSF based on the linearized Gaussian pulse that is

obtained as the principal pulse in the PAM decomposition of a binary CPM signal with modulation index 0.5. The time domain samples of the FDPSF  $q_t(n)$  can be chosen as:

$$q_t(n) = p_0(t) \Big|_{t=\tau_0 + \frac{nT}{s}}$$

where  $p_0(t)$  is the linearized Gaussian pulse,  $s$  is the over-sampling factor, and the factor  $BT$  controls the

characteristics of the waveform, and  $\tau_0$  is constant time offset, and  $n$  is an integer. Since the pulse is time limited,

$$\frac{-sM}{2} \leq n \leq \frac{sM}{2} - 1$$

the values of  $n$  can be taken in the range

With an over-sampling rate of  $s$ , the Fourier transform of  $q_t(n)$  is periodic with a period  $\frac{s}{T}$ . The FDPSF with a span of  $sM$  subcarriers is obtained by taking a  $sM$  point DFT of  $q_t(n)$  as defined in (11). Alternatively, the FDPSF can be obtained by taking an  $sM$  point DFT first as

$$\tilde{q}(m) = \sum_{n=0}^{sM-1} q_t(n) e^{-j2\pi mn/sM} \quad 0 \leq m \leq sM-1$$

Then, the left and right halves of the DFT output can be swapped so that the zero frequency components in the middle. Alternatively, the FDPSG can be obtained using a two sided DFT as

$$q(m) = \sum_{n=-\frac{sM}{2}}^{\frac{sM}{2}-1} q_t(n) e^{-j2\pi mn/sM} \quad -\frac{sM}{2} \leq m \leq \frac{sM}{2}-1$$

For the special case of  $s=1$ , one can design PDPSF without excess BW. In this case, the waveform introduces ISI but has zero multi-user interference. To obtain the time domain samples for  $s=1$ , we can first generate the samples corresponding to  $s=2$ , then choose either the even or odd symbol spaced sample sequence to generate the required FDPSF.

**Note:** The frequency domain FDPSF can be applied to a selected user who is located at the cell edge (low SNR) for increasing the transmit power. Remaining users can use conventional QAM modulation without applying FDPSF.

### Linearized GMSK Pulse

The principal pulse  $p_0(t)$  is the main pulse in Laurent's decomposition [3] is given by

$$p_0(t) = \begin{cases} \prod_{k=1}^{L_1} c(t-kT) & t \in [0, (L_1+1)T] \\ 0 & \text{otherwise} \end{cases}$$

where

$$c(t) = \begin{cases} \cos\left(-\frac{\pi}{2} q(t)\right) & t \in [0, L_1T] \\ c(-t) & t \in (-L_1T, 0] \\ 0 & |t| \geq L_1T \end{cases}$$

The pulse  $q(t)$  is a Gaussian filtered rectangular pulse response defined as

$$q(t) = \frac{1}{T} \left[ Q\left(\gamma\left(\frac{t}{T} - \frac{1}{2}\right)\right) - Q\left(\gamma\left(\frac{t}{T} + \frac{1}{2}\right)\right) \right]$$

where  $\gamma \cong \frac{2\pi BT}{\sqrt{\ln(2)}}$ ,  $BT$  is a parameter that controls the pulse shape, and  $Q(x) \cong \frac{1}{\sqrt{2\pi}} \int_x^\infty e^{-\frac{u^2}{2}} du$ . The value of  $L_1$  determines the pulse duration. Typically this value is chosen to be in the range 4-6.

## 3. Results and Discussion

In simulation, the PAPR of the waveform is measured in dB for each OFDM symbol and the cumulative distribution function (cdf) is plotted in Fig 1 without windowing. We assume  $M=48$ . The PAPR at 99 % cdf point is recorded for different waveforms. MSK corresponds to  $BT=\infty$ . Using  $s=3$ , MSK spans  $3M$  subcarriers and gives a PAPR of 0.09 dB. The GPO modulator generates a conventional MSK-like signal with approximately constant envelope. Reducing the BW occupancy of the signal to less than  $3M$  increases the PAPR to 1.95 dB for  $s=1$ . If the excess BW is restricted to 100 percent i.e., for  $s=2$ ,  $BT=0.3$  gives a PAPR of 0.79 dB that is less than of MSK with  $s=2$ . If the system does not have any excess BW i.e., for  $s=1$ ,  $BT=0.3$  Type-A pulse gives a PAPR of 1.97 dB. The PAPR reduction obtained for any of these proposed pulses is significantly less than the case of rectangular FDPSF with  $s=1$

that has a PAPR of 5.16 dB. For reference, the PAPR of QPSK modulation with rectangular FDPSF is shown in Table-1 as 6.18 dB.

In Fig 2-5, the average psd of proposed waveform for different BW allocations with and without RC window is given. It can be seen that the RC window significantly reduces the OBE compared to standard OFDM case in all cases.

For BER simulations we assumed 5 contiguous users with equal BW allocation. A WL MMSE receiver [6] is used that is able to mitigate the ISI and IUI introduced by the FDPSF. In Fig 6, the BER results for the case of rectangular FDPSF and for  $(BT=0.3, s=2)$ ,  $(BT=0.3, s=1)$  and  $(BT=\infty, s=3)$  cases is given. The results are averaged over 10000 OFDM symbol realizations. The BER of widely linear (WL) MMSE receivers for the considered cases are close to that of rectangular FDPSF. However, the conventional MMSE with  $(BT=0.3, s=2)$  has approximately 1.9 dB loss at 1% BER and for the case of  $(BT=\infty, s=3)$  the loss is 0.9 dB. We see that the WL receiver is able to fully mitigate both ISI and multi-user interference created by the pulse shaping filters. The receiver [6] uses joint detection of multiple users. We remark here that, since the IUI introduced by the FDPSF is small, conventional single user WL frequency domain equalizer can be used to obtain a similar performance.

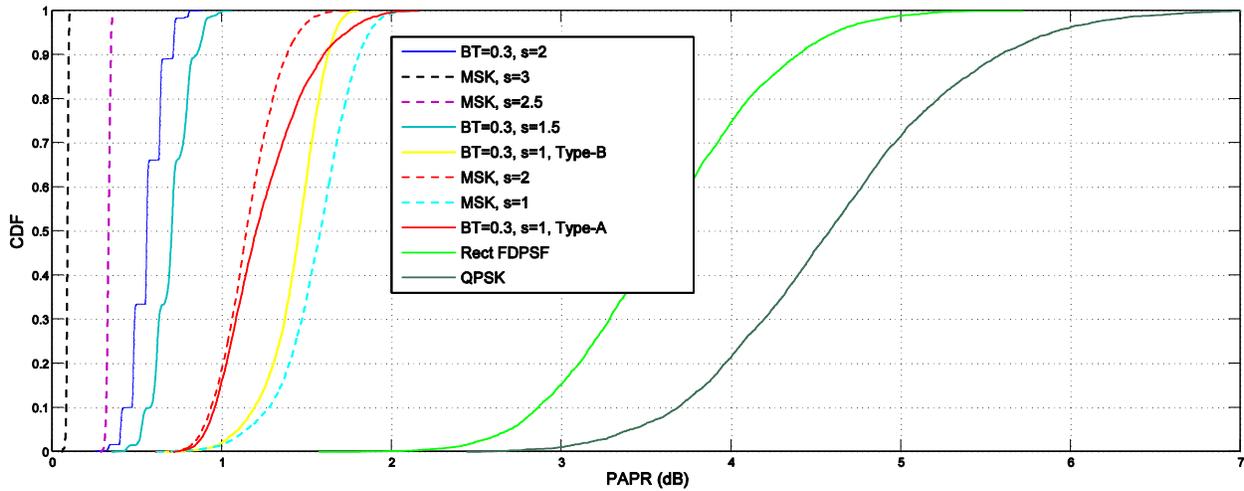


Figure1: PAPR without windowing

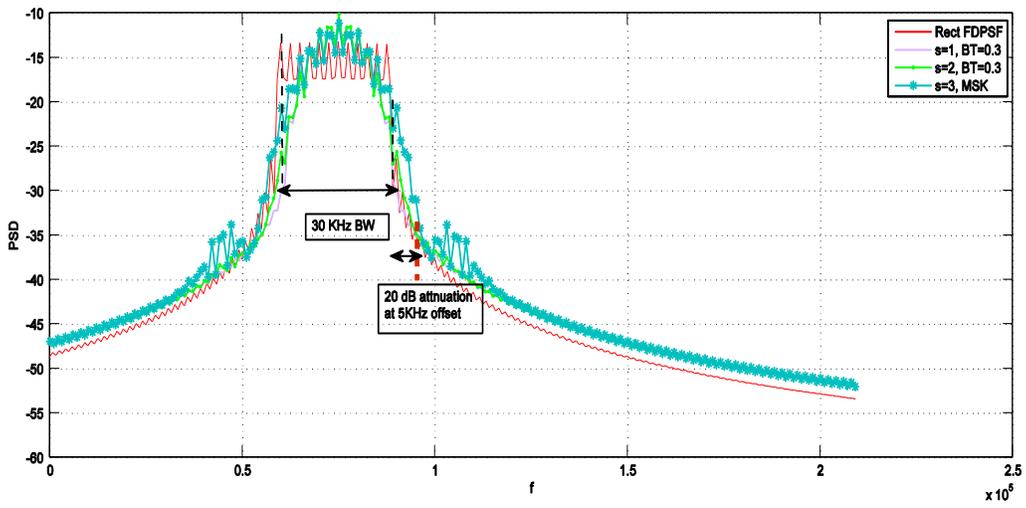


Figure2: PSD-30KHz without windowing

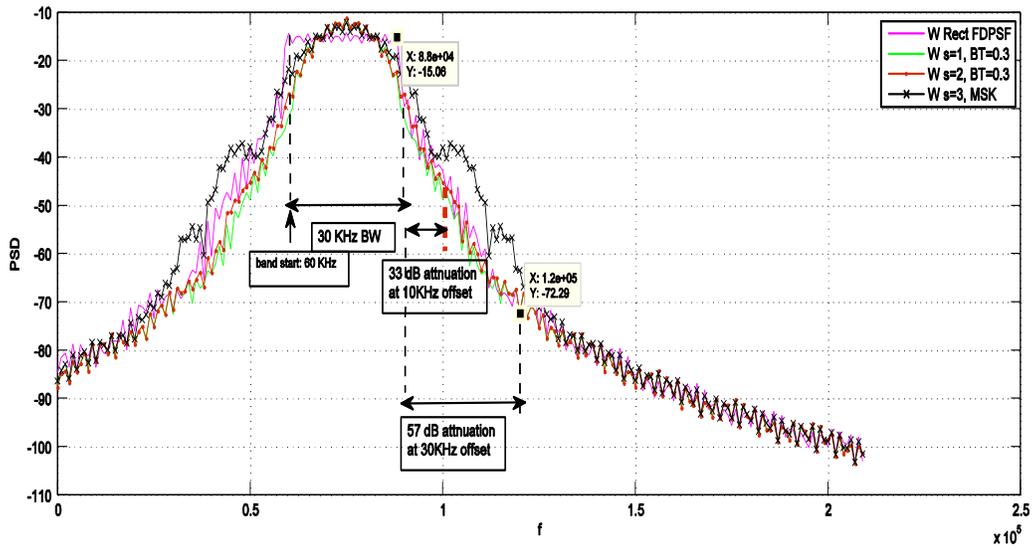


Figure3: PSD-30 KHz with windowing

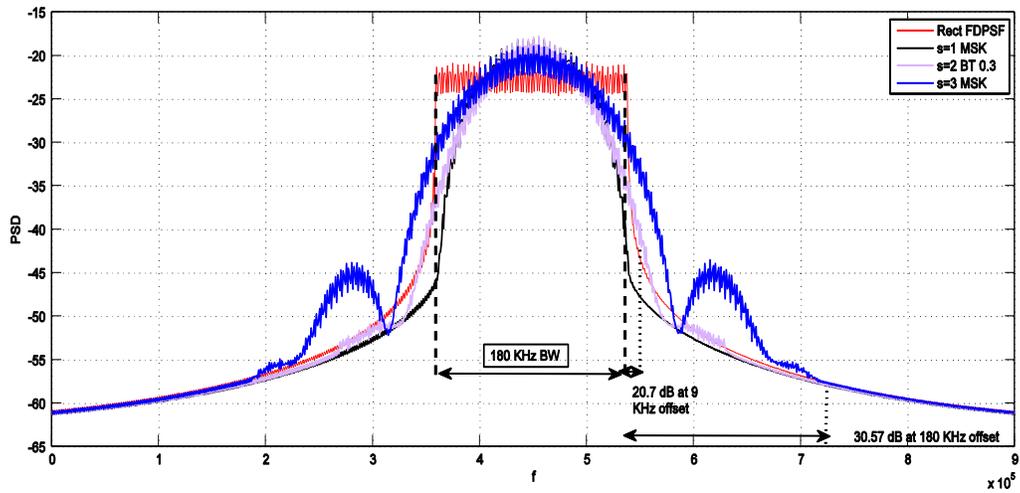


Figure4: PSD-180 KHz without windowing

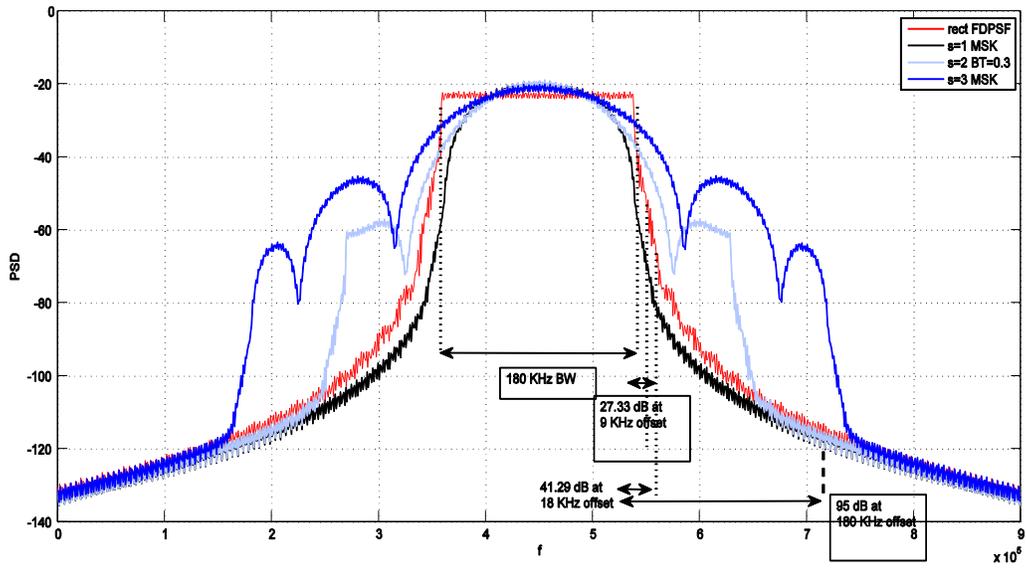


Figure5: PSD-180KHz with windowing

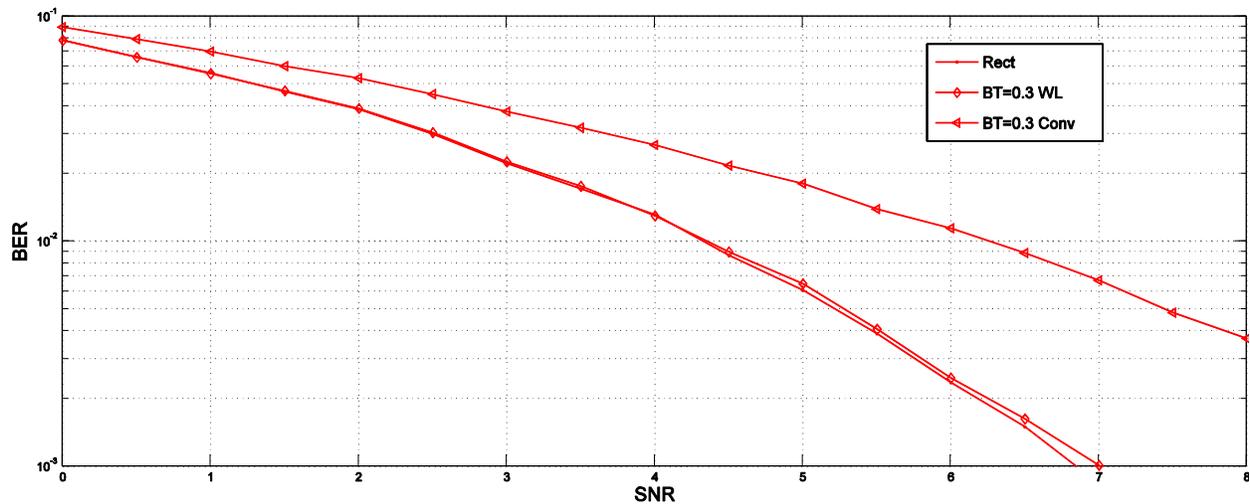


Figure6: BER

## 4. Observations

The results of this contribution shows that a) the proposed waveform is a suitable for achieving low PAPR for NB IoT. b) Further, results show that use of RC window is very effective in controlling the out of band emission. This technique can also be utilized in the downlink to reduce adjacent channel leakage.

## Proposal

It is recommended that the waveform be considered for further evaluation

## 5. References

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