On Companding and Optimization of OFDM Signals for Mitigating Impulsive Noise in Power-line Communication Systems

Kelvin Anoh*, Bamidele Adebisi*, Khaled M. Rabie*, Mohammad Hammoudeh*, and Haris Gacanin†
*School of Engineering, Manchester Metropolitan University, UK
†Nokia Bell Labs, Belgium
[k.anoh, b.adebisi, k.rabie, m.hammoudeh]@mmu.ac.uk, haris.gacanin@nokia-bell-labs.com

***Abstract*—Generally, the probability density function (PDF) of orthogonal frequency division multiplexing (OFDM) signal amplitudes follow the Rayleigh distribution, thus, it is difficult to correctly predict the existence of impulsive noise (IN) in powerline communication (PLC) systems. Compressing and expanding the amplitudes of some of these OFDM signals, usually referred to as companding, is a peak-to-average power ratio (PAPR) reduction technique that distorts the amplitudes of OFDM signals towards a uniform distribution. We suggest its application in PLC systems such as IEEE 1901 powerline standard (which uses OFDM) to reduce the impacts of IN. This is because the PLC channel picks up impulsive interference that the conventional OFDM driver cannot combat. We explore, therefore, five widely used companding schemes that convert the OFDM signal amplitude distribution to uniform distribution to avail the mitigation of IN in PLC system receivers by blanking, clipping and their hybrid (clipping-blanking). We also apply nonlinear optimization search to find the optimal mitigation thresholds and results show significant improvement in the output signal-to-noise ratio (SNR) for all companding transforms considered of up to 4 dB SNR gain. It follows that the conventional PDF leads to false IN detection which diminishes the output SNR when any of the above three nonlinear memoryless mitigation schemes is applied.**

***Index Terms*—Companding, OFDM, powerline communication (PLC), impulsive noise (IN), peak-to-average power ratio (PAPR), amplitude distribution, optimization, uniform distribution.**

I. Introduction

As the realization of internet of things (IoT) unfolds, more efforts are being channeled towards perfecting powerline communication (PLC) systems design. At homes, for example, these PLC networks can penetrate areas of poor wireless signal strengths and require no additional infrastructure hence saving cost [1]. The PLC standard, such as IEEE 1901 among others, uses the conventional electric power cables for data communication at homes and microgrids [2], [3]. It follows that PLC system can improve home automation, monitoring, security, control and comfort. However, communication data over PLC channels are garbled by impulsive noise (IN) and require optimal mitigation solutions.

The IEEE 1901 standard uses orthogonal frequency division multiplexing (OFDM) over powerlines due to its robustness over impulsive channels by applying cyclic prefix in the order of the length of the impulse response and can be implemented using fast Fourier transform (FFT) or wavelet transform [2], [4] [5], [6]. A major problem with using OFDM in PLC systems is that the asymmetric amplitude distribution gives false information about the existence of IN. Thus, applying nonlinear IN mitigation schemes namely blanking, clipping or hybrid clipping-blanking [7]–[11] realizes outputs whose signals may have been erroneously mitigated. While the asymmetrical amplitude distribution of the conventional unmodified OFDM signals leads to high peak-to-average power ratio (PAPR) problem, the erroneous IN mitigation diminishes the output signal-to-noise ratio (SNR) and consequently the bit error ratio (BER). High PAPR problem in OFDM systems lead to high power consumption of power amplifiers (PAs) and induces distortion outside the linear region of the HPA further degrading BER [12]–[15].

Knowing these, the asymmetric amplitude distribution of the PLC-OFDM systems that follow the Rayleigh distribution can be converted to a symmetrical amplitude distribution to enhance the identification, isolation and removal of the IN more effectively. In the literature, this problem has been approached by reducing the PAPR of the conventional OFDM system using partial transmit sequence (PTS) [16], selective mapping [9], constant envelope OFDM [10], iterative clipping and filtering (ICF) [17] and companding transform [11], [17]. We showed recently in [17] that companding scheme outperforms ICF in IN mitigation, although only one companding style [18] as in [11] was used. Companding OFDM signals is achieved by simultaneously compressing high amplitudes OFDM signal amplitudes and expanding the low amplitude ones toward a uniform distribution [18]–[21]. Then, when passed through an impulsive channel such as the PLC channel, the occurrence of IN can be easily identified and mitigated. While PAPR reduction schemes are applied at the transmitter, it is worthy to note that IN mitigation techniques are applied at the receiver.

The standard μ-law companding (MC) technique [18] only was studied in [11] to enhance IN mitigation without comparison to any other companding scheme including the conventional PLC-OFDM system. In this paper, we explore five companding techniques involving error-function companding (ERFC) [19], MC [18], exponential companding (EC) [20], modified log-based MC (LMC) [22] and hyperbolic arcsine companding (HASC) [23] transforms. Although [11] used MC to describe the improvement of IN mitigation at the PLC receiver, specific examples to where the optimality of performances existed were not demonstrated. Secondly, MC

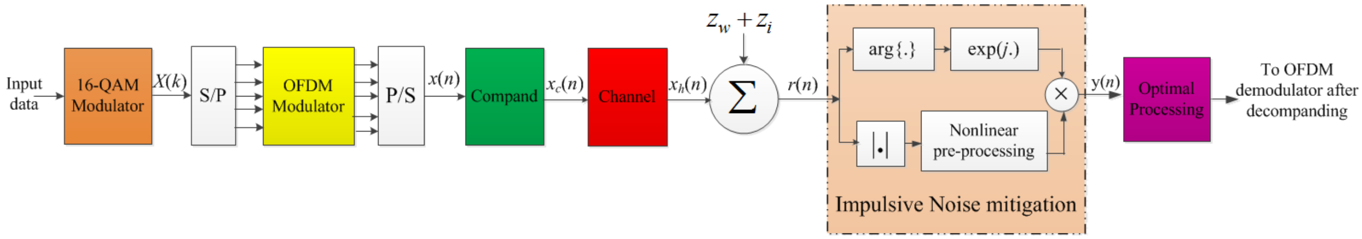

Figure 1. PLC system model with companding to enhance nonlinear IN mitigation and optimal amplitude determination before OFDM demodulation at the receiver

exists among other companding techniques and may not be the most performing. PLC systems pick up interference from other home appliances that influences the performance of transmitted signals over the powerline channels. For example, home devices induce some non-Gaussian noise into the PLC channel that corrupt the signal amplitude and consequently makes it unrecoverable. In literature, clipping, blanking and hybrid clipping-blanking are nonlinear methods used in mitigating IN [8]. Since hybrid clipping-blanking exhibits better mitigation efficiency, we use it to mitigate IN for both background noise and multipath fading channels of PLC systems.

Meanwhile, in this paper, we investigate the use of companding technique, to both improve the reduction of PAPR as well as the mitigation of IN. For example, by companding, the lower amplitude OFDM symbols are expanded while the higher amplitude OFDM symbols are compressed. This enables the entire symbol amplitudes to have a uniform distribution leading to significant PAPR reduction while making the IN identifiably clearer. Consequently, our contributions include 1) a survey/comparison of five different companding PAPR reduction techniques able to enhance OFDM signal transmission that cannot be severed by HPA; 2) converting OFDM signal amplitude distribution to uniform distribution enhances PAPR reduction and IN presence mitigation using nonlinear pre-processing; 3) probability distribution function (PDF) of PAPR reduction styles achieving OFDM amplitude distribution closer to uniform distribution dispenses with higher output SNR due to increased IN mitigation efficiency than unmodified OFDM amplitude distribution; 4) reducing PAPR of OFDM signal before transmission uncovers some IN presence masqueraded in the unmodified amplitude distribution; 5) we search for the optimal blanking/clipping threshold and find that optimal nonlinear pre-processing amplitude thresholds are far below the ones presented by the unmodified OFDM amplitude distribution system. In total, our results will enhance the design of energy-efficient and high-throughput PLC communication systems. For example, with PAPR reduction the output SNR at high IN probability attains 1.4dB and 1.7dB at low IN probability better than unmodified OFDM system. With optimal search, the clipping/blanking threshold reduces to 1.4 and 1.3 respectively for low and high IN probabilities increasing the output SNRs to 4dB and 2.6dB.

The remaining parts of this paper are organized as follow. The system model is described in Section II including the different companding styles showing model companding performance, PDF and PAPR performances. In Section VI, the performances of the five companding models under investigation are presented and discussed in terms of PDF, PAPR and SNR. The conclusion follows in Section VII.

II. System Model

The general system model considered in this study is illustrated in Fig. 1. In the system model, we consider an OFDM-driven powerline system that picks up IN as it traverses the cable. Numerically, consider a frequency domain data symbol $X = [X_0, X_1, X_2, \cdots, X_{N-1}]$ which can be converted into its time domain component by passing the signal through an IFFT-block as

$$x(n) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X(k) \exp\left(j2\pi \frac{nk}{N}\right) \forall n = 0, 1, \cdots, N-1. \tag{1}$$

When separated into its component parts, namely real (x_r) and imaginary (x_i), the time domain signal in (1) can be characterized from the knowledge of central limit theorem. For example, x_r and x_i are identically and independently distributed (i.i.d.) Gaussian random variables. It follows that if $x(n)$ is sufficiently large, the PDF of these x_r and x_i follows a Rayleigh distribution. In other words, if $x \sim \mathcal{N}\left(\mu_x, \sigma_x^2\right)$, then

$$f_{|x|}(x; \mu_x, \sigma_x) = \frac{1}{\sqrt{2\pi\sigma_x^2}} \exp\left(-\frac{1}{2}\left(\frac{x_0 - \mu_x}{\sigma_x}\right)^2\right) \tag{2}$$

where $f_{|x|}(x; \mu_x, \sigma_x)$ is the PDF, $\mu_x = \mathbb{E}\{x(n)\}$, $\mathbb{E}\{\cdot\}$ is the statistical expected mean operator, σ_x is the standard deviation, $\sigma_x^2 = \frac{1}{2}\mathbb{E}\left\{|x(n)|^2\right\} = 1$ is the variance of $x(n)$ whose discrete envelope is x_0 and $|\cdot|$ computes the absolute value of the input variable. When $x(n)$ is normally distributed, then $\mu_x = 0$. It is well-known that the cumulative distribution function (CDF) of the Gaussian distributed variable $x(n)$ is the integration of the PDF. Thus, the CDF of (2) becomes

$$C_x\left(x; \mu_x, \sigma_x\right) = \frac{1}{2}\left(1 + \text{erf}\left(\frac{1}{\sqrt{2}}\left(\frac{x_0 - \mu_x}{\sigma_x^2}\right)\right)\right) \tag{3}$$

where $\text{erf}\left(z\right) = \frac{1}{\pi}\int_{-z}^{z} e^{-u^2} du$. Pictorially, the PDF expression in (2) for large number of OFDM signal subcarriers $N = 4096$ can be represented in Fig. 2. Clearly, the picture shows the Rayleigh distributed OFDM signal amplitudes. A majority of the signals have amplitudes distributed around the μ_x while a few others are distributed below and above the mean amplitude of the distribution. The amplitudes distributed above μ_x right-hand side lead to high PAPR in the

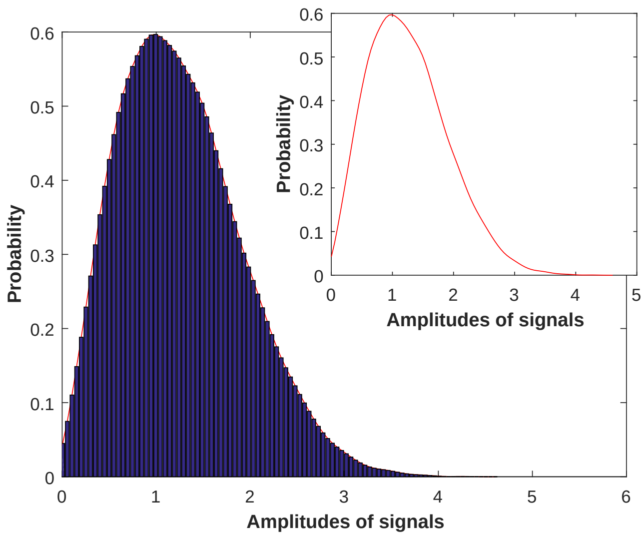

Figure 2. PDF distribution of conventional OFDM signal amplitudes which gives information into its possible behaviour over powerline channel

conventional OFDM. In addition, the fraction of amplitudes dominating the upper bound of the distribution hides the IN and can be erroneously clipped or blanked during the nonlinear IN mitigation; this will diminish the output SNR and BER performances.

If the amplitude distribution of the OFDM signal is flat, then the mitigation of IN can be easier and the performance of the system will be improved. Our goal therefore is to apply transforms that can convert the amplitude distribution of the conventional OFDM system to enhance the identification and mitigation of IN in powerline systems. Although this can be achieved by either clipping or companding, we appeal to companding due to its light-weight on the system.

III. Methods of Reducing PAPR Approximating the PDFs of OFDM to Uniform Distribution by Companding Transforms

As a rule of the thumb, companding transforms can be realized by deriving a suitable PDF and then the CDF that can reduce PAPR of OFDM systems. As an example, recall the time-domain OFDM signals in (1), the companding transform uses the following identity [19], [21]

$$\mathcal{F}(\boldsymbol{x}(n)) = C_{x_c}^{-1}\left(C_x\left(x\left(n\right)\right)\right) \tag{4}$$

to convert the PDF to a suitable PAPR reduction model, where $C_x(\cdot)$ is the CDF of the uncompanded signal, $C_{x_c}(\cdot)$ and $C_{x_c}^{-1}(\cdot)$ is the CDF and inverse CDF of the CDF of companded signal respectively. Then resulting model $\mathcal{F}(\boldsymbol{x}(n))$ is the required companding transform that converts the PDF of conventional OFDM to a desired distribution. Since $|x_c(n)|$ tends to the desired uniform distribution, then the CDF is written as [19]

$$C_{x_c(n)}\left(x_c\right) = \frac{x_c}{2A} + \frac{1}{2},\ 0 \le x_c \le A. \tag{5}$$

Meanwhile, by combining (3), (4) and (5), the companding transform can be expressed as

$$\mathcal{F}_1(\boldsymbol{x}(n)) = \text{sgn}\left(x\right)\cdot A_1 \cdot \text{erf}\left(\frac{|x|}{\sqrt{2\sigma_x^2}}\right),\ 0 \le x \le 1 \tag{6}$$

where $A_1 = \sqrt{3\sigma_x^2}$. Our target is however to achieve $[0,\ A]$, where $0 < |\mathcal{F}(\boldsymbol{x}(n))| \le 1$. Another example can be achieved from the well-known mu-law of the form [18]

$$\mathcal{F}_2(\boldsymbol{x}(n)) = A_2\,\text{sgn}\left(x(n)\right)\frac{\ln\left[1+\mu\left|\frac{x(n)}{A_2}\right|\right]}{\ln\left(1+\mu\right)} \tag{7}$$

where A_2 is a normalization parameter confined within $0 \le \left|\frac{x(n)}{A_2}\right| \le 1$. The problem with (7) is that it expands the amplitudes of lower energy signals without compressing the larger ones. Thus, in [22], (7) was modified as

$$\mathcal{F}_3(x(n)) = \text{sgn}(x)\left(\alpha_3 \times \ln\left[1+\mu\left|\frac{x(n)}{A_3}\right|\right]\right)^{\frac{1}{2a_3}} \tag{8a}$$

where

$$\alpha_3 = \left(\mathbb{E}\left\{|x|^2\right\}/\mathbb{E}\left\{\sqrt[a_3]{\left(\log\left(1+\mu\frac{|x|}{A_3}\right)\right)^{b_3}}\right\}\right). \tag{8b}$$

Based on the hyperbolic arcsine function which can also be expressed as $\log\left(x+\sqrt{x^2+1}\right) = \sinh^{-1}(x)$, the authors in [23] suggested a different companding transform of the form

$$\mathcal{F}_4(x(n)) = \beta_4 \times \begin{cases}\text{sgn}(x)\times\sinh^{-1}\left(K\left|x\right|\right), & |x| \le cA_4\\ \text{sgn}(x)\times\sinh^{-1}\left(KcA_4\right), & |x| > cA_4\end{cases} \tag{9}$$

where c and K are the flexing-point determining parameters associated with A_4. β_4 is the parameter that normalizes the output power of the companded signal to be similar to that of the input signal. In [20], an exponential companding transform that can approximate the PDF of the conventional OFDM signal to a uniform distribution was proposed such as

$$\mathcal{F}_5(x(n)) = \text{sgn}\left(x\right)\sqrt[d_5]{\beta_5\left[1-\exp\left(-\frac{x^2}{\sigma^2}\right)\right]} \tag{10a}$$

$$\beta_5 = \left(\frac{E\left\{|x(n)|^2\right\}}{E\left\{\sqrt[d_5]{\left[1-\exp\left(-\frac{x^2}{\sigma^2}\right)\right]^2}\right\}}\right)^{\frac{d_5}{2}} \tag{10b}$$

where $d_5 > 0$ and in general, $\text{sgn}\left(x\right) = \frac{x(n)}{|x(n)|}$ is the phase. In Fig. 3, we analyze the companding performances of these transforms in terms of amplitude compression and expansion respectively of input signals. MC scheme expands the amplitudes of lower energy signals without impacts on the high energy signals. This will increase the output SNR. It is followed by LMC although LMC expands the amplitude of lower energy signals and also compresses the amplitudes of higher energy signals; this is similarly true for ERFC. On the other hand, HASC does not impact the amplitudes of lower energy signals, however it compresses the amplitude of high energy signals. Lastly, the EC compresses the amplitudes of high amplitude signals and expands the low energy ones;

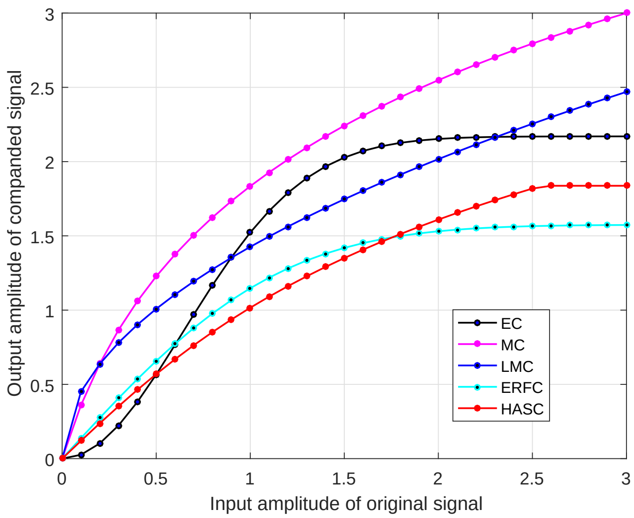

Figure 3. Comparisons of different companding transforms for converting Rayleigh PDF to a uniform distribution; $d_5 = 8$, $\mu_{LMC} = 256$, $\mu_{MC} = 256$, $c = 0.85$, $K = 1.8$, $a_3 = 2$ and $b_3 = 2$,

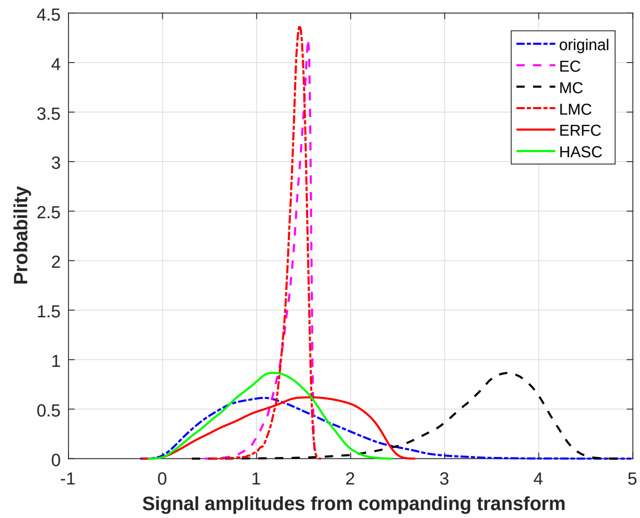

Figure 4. Characteristic PDFs of original and companded signal informing on the amplitude distributions of the signal, $d_5 = 8$, $\mu_{LMC} = 256$, $\mu_{MC} = 256$, $a_3 = 2$ and $b_3 = 2$

this will greatly improve the distribution towards the desired uniform distribution. Meanwhile, it is worthy to mention that companding involves compressing the amplitudes of the amplitudes of large amplitude (high energy) signals and expanding low amplitude (low energy) signals. While some companding transforms achieve the former, others achieve the latter only. Ideally, effective companding transforms achieve both characteristics simultaneously and their performances in availing IN mitigation lies on how much of uniform distribution that is achieved.

In general, the result of the output companded signal can now be expressed as

$$x_{ct}(n) = \mathcal{F}\left(\boldsymbol{x}\left(n\right)\right), \ \forall n = 0, 1, \cdots, N-1. \tag{11}$$

While some companding schemes either expands lower energy contents of $|x_{ct}(n)|$ only (for example, MC), others compress higher energy contents of $|x_{ct}(n)|$ only such as HASC. The former unfairly increases the $x_c(n)$ at the expense of the original signal (when compared) while the latter decreases the energy of $x_c(n)$ which will lead to poor SNR performance. Consequently, to ensure that both the output companded signal and original signal dispenses with comparable energy, we scale the energy in $x_{ct}(n)$ using that of $x(n)$ as

$$\beta = \sqrt{\left\{\frac{\mathbb{E}\left\{|\boldsymbol{x}(n)|^2\right\}}{\mathbb{E}\left\{|\boldsymbol{x}_c(n)|^2\right\}}\right\}}. \tag{12}$$

Thus, $x_{ct}(n)$ is scaled by β prior to further processing or transmission. When OFDM signals are companded, the signal undergoes some nonlinear amplitude distortion and can be described from the Bussgang theorem as [24]

$$x_{cc}(n) = \mathcal{F}\left(\boldsymbol{x}\left(n\right)\right) = \alpha_n x(n) + D_n \ \forall n = 0, 1, \cdots, N-1 \tag{13}$$

where α_n is the attenuation factor and D_n is distortion noise. Thus, before transmission, we compensate the amplitude distorted signal as follows

$$x_c(n) = \bar{\alpha}_n x_{cc}(n), \tag{14}$$

where $\bar{\alpha}_n$ is the correlation coefficient of the distorted and original signal that minimizes the error in $\mathbb{E}\left[|\boldsymbol{x}(n) - \mathcal{R}x^*_{ct}(n)|^2\right]$ after power amplification/reduction from companding which can be written as [13]

$$\bar{\alpha}_n = \frac{\mathbb{E}\left[\boldsymbol{x}(n)\, x^*_c(n)\right]}{\mathbb{E}\left[|x_c(n)|^2\right]}, \tag{15}$$

where $(\cdot)^*$ represents complex conjugate operator and $\bar{\alpha}_n = \mathcal{R}$.

A. Characteristic PDFs of the Companded signal output

We explore the characteristic PDFs (Fig. 4) of the companding transforms which provides information into the PAPR reduction characteristics of the transform and consequently the ability to influence the identification, isolation and mitigation (removal) of the IN present in the transmitted signal.

From Fig. 4, the EC scheme converts almost all the signal amplitudes to 1.5, in other words towards a perfect uniform distribution. LMC on other hands converts these amplitudes with almost perfect uniform distribution. The rest schemes smear the amplitude distributions uniformly. Notice that MC expands the amplitudes of low energy signals; this will lead to an undue high output SNR. Then by (12), we apply the normalization parameter to ensure equal power dissipation with original input signal before signal transmission, which also affects the PDF of the MC scheme as shown in Fig. 5. We observe that the PDF of the MC scheme improves towards a better uniform distribution by scaling the distribution with β. We shall discuss the respective PAPR performances based

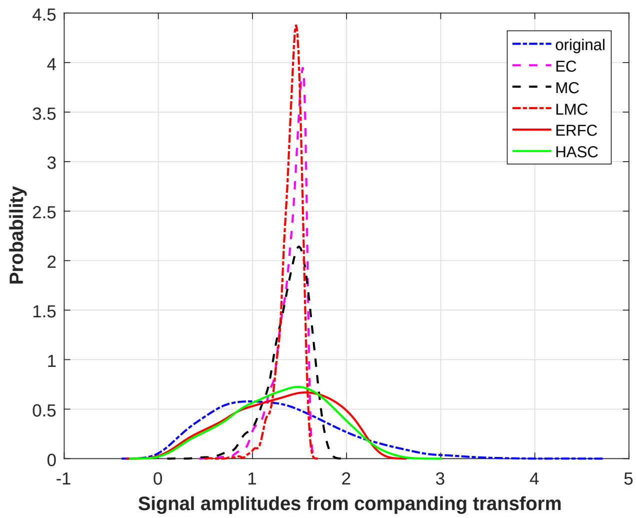

Figure 5. Characteristic PDFs of original and companded signal informing on the amplitude distributions of the signal (with β), $d_5 = 8$, $\mu_{LMC} = 256$, $\mu_{MC} = 256$, $a_3 = 2$ and $b_3 = 2$

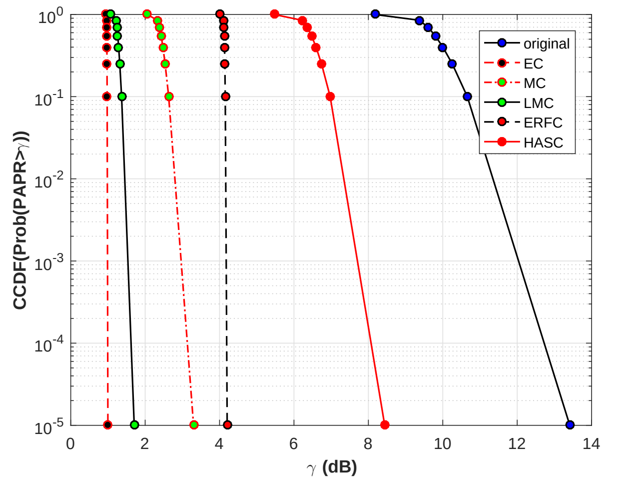

Figure 6. PAPR comparisons of different companding transforms, $d_5 = 8$, $\mu_{LMC} = 256$, $\mu_{MC} = 256$, $a_3 = 2$ and $b_3 = 2$, $\ell = 4$

of the resulting amplitude distributions realized from these companding transform reflecting how the proximity to uniformity improves the PAPR performances of the respectively driven OFDM systems in Section III-B. This is the desired performance for IN mitigation and will be shown shortly in Section VI.

B. Performance evaluation of the PAPR reduction levels of the companding schemes

The PAPR of OFDM system informed by the excess amplitudes above the mean amplitude gives information into the resulting distribution of the signal. From the foregoing discussion, we now know that distorting the amplitudes towards a uniform distribution will impact the impulsive mitigation. In other words, as$|x(n)| \to 1$ then the PAPR$(x) \to 0$dB then IN can be easily identified. Consequently, let the PAPR of an undistorted OFDM signal frame be [25]

$$\text{PAPR}\left(x\left(n\right)\right) = 10\log_{10}\left\{\frac{\max\limits_{n=0,1,\cdots,\ell N}\left(\left|x(n)\right|^2\right)}{\frac{1}{\ell N}\sum\limits_{n=0}^{\ell N-1}\left(\left|x(n)\right|^2\right)}\right\} \quad (16)$$

where ℓ is oversampling factor. The complementary CDF (CCDF) of $x(n)$, namely $CC_X = 1 - C_X$, is used to measure the performance of PAPR [22], [26], where C_X is the CDF described as

$$C_X = \Pr\left\{|\boldsymbol{x}(n)| \leq \gamma\right\} = 1 - \exp\left(-\frac{x_0^2}{\sigma_x^2}\right), \ \forall x_0 \geq 0 \quad (17)$$

where x_0 is the discrete envelope of $x(n)$. In other words, C_X measures the probability that the amplitude of the current OFDM signal exceeds a target threshold, γ. Considering all points, i.e., $n = 0, 1, \cdots, \ell N - 1$, the CCDF can be rewritten as CC

$$CC_X = \Pr\left\{|x(n)| > \gamma\right\} = \left[1 - \left(1 - \exp\left(-\frac{x_0^2}{\sigma_x^2}\right)\right)^{\ell N}\right]. \quad (18)$$

We can similarly measure the PAPR of the companded signal using CCDF. For example, the PAPR of the companded signals can be described (from the CCDF) as

$$CC_{x_c} = \Pr\left\{|\mathcal{F}(\boldsymbol{x}(n))| > \gamma\right\} = \Pr\left\{|\boldsymbol{x}_c(n)| > \gamma\right\}. \quad (19)$$

Ideally, a standard companding function, $\mathcal{F}(\cdot)$, increases the amplitude of the smaller signals and compresses the higher amplitudes towards 1. However, some companding transforms either reduce the amplitudes of the higher symbols or increase the amplitudes of the lower symbols only. At the receiver, the decompanded signal is given by $\hat{\boldsymbol{x}}\left(n\right) = \mathcal{F}^{-1}\left(\mathcal{F}\left(\boldsymbol{x}\left(n\right)\right)\right)$. Furthermore, in Fig. 6, we consider the performances of the PAPRs for the different companding techniques.

While all the companding transforms reduce the PAPR of the original uncompanded OFDM signals, some transforms perform better than others. We can link the performances of the PAPR reduction schemes to the PDF being verily uniformly distributed in Fig. 4. As the most uniformly distributed, EC technique achieves the best PAPR reduction as depicted in Fig 6. Secondly, the LMC and MC follow the EC in performance, respectively, which can be improved by varying μ. These results are important for the IN identification and mitigation.

IV. Transmission of Companded OFDM Signal over Powerline Channels with Impulsive Noise

From the foregoing discussion, it can be inferred that distorting the conventional PDF of an unmodified OFDM system can change the PAPR performance. Then, converting the PDF to a uniform distribution hence avails the presence of IN better. Consequently, considering the transmission of OFDM signal over a memoryless IN channel with characteristic Gaussian noise, $z_w \sim \mathcal{N}\left(0, \sigma_w^2\right)$ and IN $z_i \sim \mathcal{N}\left(0, \sigma_i^2\right)$ where σ_w^2 and σ_i^2 are the variances. Let the Gaussian and the IN samples represented as $z(n) = z_w(n) + z_i(n)$ be uncorrelated, then the distribution fits into the mixture-Gaussian model with characteristic PDF as follows [8], [27]

$$f_z\left(z; \mu_z, \sigma_z\right) = \sum_{l=0}^{L=1} p_l \mathcal{N}\left(z_0(n); 0, \sigma_{z,l}^2\right) \forall n = 1, 2, \cdots, N-1 \tag{20}$$

where $\mathcal{N}\left(z_0; 0, \sigma_{z,l}^2\right) = \frac{1}{\sigma_{z,l}\sqrt{2\pi}} \exp\left(-\frac{1}{2}\left(\frac{z_0-\mu_z}{\sigma_{z,l}}\right)^2\right)$ is the Gaussian PDF of $z(n)$ with z_0 discrete envelope, zero-mean ($\mu_z = 0$), variance $\sigma_{z,l}^2$ and p_l is the mixing probability of the l^{th} noise component. From (20), we can separate the mixing probability into $p_0 = 1 - p$ and $p_1 = p$, where p is the probability of IN occurrence. Similarly, the variance can be separated into $\sigma_{z,0}^2 = \sigma_w^2$ and $\sigma_{z,1}^2 = \sigma_w^2 + \sigma_i^2$.

In general, therefore, the received signal at the destination PLC modem can be expressed as

$$\begin{aligned} \boldsymbol{r}(n) &= \boldsymbol{x}_c(n) + \boldsymbol{z}(n) \\ &= \mathcal{F}(\boldsymbol{x}(n)) + \boldsymbol{z}_w(n) + \boldsymbol{z}_i(n). \end{aligned} \tag{21}$$

Given the signal power $\sigma_{x_c}^2$ of the companded signal, the input SNR and signal-to-interference noise ratio (SINR) respective to the AWGN and IN can be expressed as

$$\chi = \left(\frac{\sigma_{x_c}^2}{\sigma_w^2}\right) \tag{22}$$

$$\psi = \left(\frac{\sigma_{x_c}^2}{\sigma_i^2}\right) \tag{23}$$

where $\text{SNR(dB)} = 10\log_{10}(\chi)$ and $\text{SINR(dB)} = 10\log_{10}(\psi)$. The IN $z_i(n)$ is non-Gaussian and thus can be expressed as [28]

$$z_i(n) = b(n) \cdot \boldsymbol{n}_w(n), \forall n = 0, 1, \cdots, N-1 \tag{24}$$

where $n_w(n) \sim \mathcal{N}\left(0, \frac{N_0}{2}\Gamma\right)$, $\forall n = 0, 1, \cdots, N-1$ with N_0 as the single-sided power spectral density and Γ is the mean power ratio of the IN and AWGN components. Meanwhile, $b(n)$ is the Bernoulli process with probability mass function defined as [29]

$$\Pr\{b(n)\} = \begin{cases} p, & b(n) = 1 \\ 1-p & b(n) = 0 \end{cases}, \forall n = 0, 1, \cdots, N-1 \tag{25}$$

Since the subcarrier frequencies of OFDM are usually narrow, the symbol durations are usually long which can mitigate the impulsive nature of the channel. This is only consistent if the impulsive energy is moderate [8]. Otherwise, when the noise amplitude is too large, one of the most effective methods is by applying some memoryless nonlinearities before the OFDM demodulator [8]. Three popular styles include clipping, blanking or a hybrid blanking-clipping [30]. In this study, we consider the hybrid clipping and blanking nonlinear preprocessing to eliminate the IN impacts on the system which achieves the best performance when considering the output signal-to-noise ratio (SNR) [8]. Then, at the receiver, the decompanded signal is given by

$$\hat{\boldsymbol{x}}(n) = \mathcal{F}^{-1}\left(x_c(n)\right) = \mathcal{F}^{-1}\left(\mathcal{F}\left(\boldsymbol{x}(n)\right)\right) \tag{26}$$

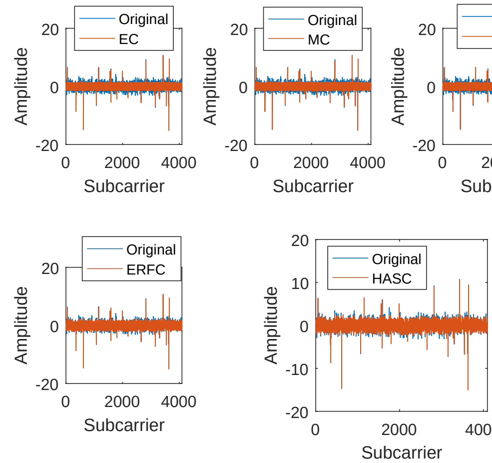

Figure 7. Amplitude behaviour of OFDM companded signals over IN channels ($p = 0.01$)

where $\mathcal{F}^{-1}(\cdot)$ is the decompanding transform. While $\mathcal{F}(\cdot)$ distorts the amplitude of OFDM signals, the bit error ratio performance of the received signal after traversing a fading channel depends (among other things) on the ability of $\mathcal{F}^{-1}(\cdot)$ to correctly restore the companded signal when presented at the receiver.

A. Amplitudes of Companded and Non-companded Symbols

Passing the companded signal through a mixture Gaussian and impulsive channel, we depict in Fig. 7 that companding OFDM signals helps to identify availing proper mitigation using memoryless nonlinear preprocessing. Comparing with the original amplitudes, the IN are notably differentiable from the OFDM signals after companding. However, using the companding transform over the original signals before the IN channel will lead to uniform distribution of the OFDM symbol thus making the IN appear with higher energy.

B. Hybrid Clipping and Blanking Nonlinear Preprocessors

We assume that the nonlinear mitigation schemes that are applied to reduce the IN effect corresponds to the signal sampled at Nyquist sampling rate. This implies that all distortion suffered by the signal are all within the in-band. For our system, the three nonlinear mitigation approaches are applied at the front-end of the OFDM system, in which all schemes are based on the ideal that for $|x_c(n)| > T$, then the corresponding signal is clipped or blanked or both applied simultaneously [8], [28], [31]. Considering clipping first, we express the output clipped signals as

$$y(n) = \begin{cases} r(n) & |r(n)| \leq T_c \\ T_c \exp\left(j \arg\{r(n)\}\right) & |r(n)| > T_c \end{cases} \tag{27}$$

where T_c is the clipping threshold. Similarly for blanking scheme, we express the output blanked signal as

$$y(n) = \begin{cases} r(n) & |r(n)| \leq T_b \\ 0 & |r(n)| > T_b \end{cases} \tag{28}$$

where $\arg\{\cdot\}$ estimates the angle and T_b is the blanking threshold. Recently, we have used blanking scheme to show that companding scheme outperforms iterative clipping and filtering PAPR reduction technique in using these to avail IN mitigation [17]. We will show in Section VI-A that combing clipping and blanking achieves better output SNR. This hybrid technique is usually expressed as [8], [32]

$$y(n) = \begin{cases} r(n), & |r(n)| \leq T_1 \\ T_1 \times \exp\left(j \arg\{r(n)\}\right), & T_1 < |r(n)| \leq T_2 \\ 0, & |r(n)| > T_2 \end{cases} \tag{29}$$

where T_2 is related to T_1 as $T_2 = 1.4T_1$ and $\arg\{r(n)\}$ computes the phase of the input signal $r(n)$. Clearly, (29) shows that the nonlinear IN mitigation schemes only operate on the amplitudes without modifying the phase. The output performance on the threshold amplitude will be determined by the degree of uniformity achieved by the companding transform.

C. Estimate of the output SNR after companding

Both companding at the transmitter and nonlinear preprocessing of the received signal at the receiver are nonlinear processes. The two schemes induce some degree of amplitude distorting noise to the system as they attempt to improve the system performance. According to Bussgang theory, we have noted in (13) that any nonlinearly distorted signal requires a scaling to the compensate for the amplitude distortion. Thus, to estimate the received SNR after the nonlinear memoryless preprocessing, we follow [8] to attain

$$SNR_{out} = 10\log_{10}\left\{\frac{\mathbb{E}\left\{|K_0\mathcal{F}(x(n)|^2\right\}}{\mathbb{E}\left\{|y(n) - K_0\mathcal{F}(x(n)|^2\right\}}\right\} \tag{30a}$$

$$= 10\log_{10}\left\{\left(\frac{E_{out}}{2K_0^2} - 1\right)^{-1}\right\} \tag{30b}$$

where $E_{out} = \mathbb{E}\left\{|y(n)|^2\right\}$ is the output power of the nonlinearly mitigated signal, K_0 is expressed in (31) and $\mathbb{E}\{\cdot\}$ is the expected value operator. where $E_{out} = \mathbb{E}\left\{|y(n)|^2\right\}$ is the output power of the nonlinearly mitigated signal, K_0 is a scaling factor which can be defined respectively as in (31) shown at the top of the next page, where $Q(s) = \frac{1}{\sqrt{2\pi}}\int_s^\infty \exp\left(-\frac{\nu^2}{2}\right)d\nu$.

V. Optimization of Preprocessing Parameters

Considering a linearly increasing amplitude thresholds, there exists an optimum for which the output SNR is maximized. Companding increases the likelihood of correctly identifying OFDM subcarrier indices corrupted by IN. We undertake the campaign of optimizing the companded OFDM system model based on the kernel PDFs of the deployed companding schemes. To start with, consider the conventional optimization model for determining the optimal output SNR as [11], [33]

$$T_{opt}^{clip/blank} = \underset{0\leq T\leq A_{max}}{\arg\max}\ \gamma^{clip/blank}\left(T, p, \psi, \chi\right) \tag{32a}$$

subject to

$$A_{max} = \arg\max_{0\leq n\leq N-1}\left(|x(n)|\right) \tag{32b}$$

where p is the probability of IN occurrence from the foregoing discussion. Recall that one of our goals, for example, is to establish the best performing companding transform that can achieve most reduction of the IN in powerline system dispensing with OFDM modulation, we apply optimization to finding the maximal received SNR and the optimal blanking threshold given the different companding transforms. Consequently, we modify (32) to include $\mathcal{F}_m(\boldsymbol{x}(n))$ as

$$T_{opt}^{clip/blank} = \underset{0\leq T\leq A_{max}}{\arg\max}\ \gamma^{clip/blank}\left(T, \mathcal{F}_m(\boldsymbol{x}(n)), p, \psi, \chi\right) \tag{33a}$$

subject to

$$A_{max} = \arg\max_{0\leq n\leq N-1}\left(|x(n)|\right) \tag{33b}$$

$$\forall m = 1, \cdots, 5$$

From (33), there exists an optimal amplitude within the companded signals for which the output SNR is optimum. Thus, the expression (33) holds the objective variable $\mathcal{F}_m(\boldsymbol{x}(n))$ which is considered one at a time. Following (30), we define $\gamma^{clip/blank}$, the output SNR after deploying the hybrid clipping-blanking nonlinear memoryless processing as

$$\gamma^{clip/blank} = \frac{\mathbb{E}\left\{K_0^{clip/blank}\left|x_c(n)\right|^2\right\}}{\mathbb{E}\left\{|y(n)|^2 - K_0^{clip/blank}\left|x_c(n)\right|^2\right\}} \tag{34}$$

where $K_0^{clip/blank}$ is a scaling parameter described in (31) that compensates the nonlinear memoryless preprocessing. Meanwhile, p_0 and $p_1$are dependent on p which have chosen to be $p = \{0.1,\ 0.01\}$ in all our investigations.

A. Transmission over Multipath Fading Channel

PLC systems characteristically exhibit frequency selectivity, frequency dependent attenuation and IN [34], [35]. With FFT, the IN is spread across the bandwidth while channel attenuation degrades the signal power. Fading in PLC channel can be modeled as multiple reflections traversing many channels routes for any selected point of interest [34]. Different fading channel models exist in the literature for PLC systems [29], [34]–[37]. However, we follow log-normal model which has the following PDF [29]

$$f_u(u;\mu_u,\sigma_u) = \frac{\zeta}{u\sqrt{2\pi\sigma_u^2}}\exp\left(-\left(\frac{10\log_{10}(u) - \mu_u}{\sqrt{2\sigma_u^2}}\right)^2\right) \tag{35}$$

where $u = h^2$, h is the channel impulse response, $\zeta = 10/\ln(10)$ is a scaling constant, σ_u^2 and μ_u are the variance and mean of $10\log_{10}(h)$, respectively. In this case, the

$$K_0^{clip} = 1 - \sum_{l=0}^{L=1} p_l \left(\mathrm{e}^{-\left(\frac{T_1^2}{2\left(\sigma_l^2+1\right)}\right)} - \sqrt{\frac{\pi}{2}} \frac{T_1}{\sqrt{\sigma_l^2+1}} Q\left(\frac{T_1}{\sqrt{\sigma_l^2+1}}\right)\right),\ K_0^{blank} = 1 - \sum_{l=0}^{L=1} p_l \left(\frac{T_1^2}{2\left(\sigma_l^2+1\right)} + 1\right) \mathrm{e}^{-\left(\frac{T_1^2}{2\left(\sigma_l^2+1\right)}\right)} \tag{31a}$$

$$K_0^{clip/blank} = \sum_{l=0}^{L=1} p_l \left(\frac{T_1^2 T_2^2}{2\left(\sigma_l^2+1\right)} \mathrm{e}^{-\left(\frac{T_2^2}{2\left(\sigma_l^2+1\right)}\right)} + \mathrm{e}^{-\left(\frac{T_1^2}{2\left(\sigma_l^2+1\right)}\right)}\right) - \sum_{l=0}^{L=1} p_l \sqrt{\frac{\pi}{2}} \frac{T_1}{\sqrt{\sigma_l^2+1}} \left[Q\left(\frac{T_1}{\sqrt{\sigma_l^2+1}}\right) - Q\left(\frac{T_2}{\sqrt{\sigma_l^2+1}}\right)\right] \tag{31b}$$

received signal involves passing the PAPR reduced signal over multipath fading channel and can be expressed as

$$\begin{aligned} \boldsymbol{r}_{chan}(n) &= h(n)\boldsymbol{x}_c(n) + \boldsymbol{z}(n) \\ &= h(n)\mathcal{F}(\boldsymbol{x}(n)) + \boldsymbol{z}_w(n) + \boldsymbol{z}_i(n). \end{aligned} \tag{36}$$

Although the IN mitigation of the received signal in (36) proceeds as illustrated in (29), the fading channel coefficient attenuates signal amplitude and thus degrades the received signal power. From (36), the output SNR after the foregoing hybrid clipping-blanking IN mitigation becomes

$$\gamma_{chan}^{clip/blank} = \frac{\mathbb{E}\left\{K_0^{clip/blank}\left|h(n)x_c(n)\right|^2\right\}}{\mathbb{E}\left\{\left|y_{chan}(n)\right|^2 - K_0^{clip/blank}\left|h(n)x_c(n)\right|^2\right\}} \tag{37}$$

where $y_{chan}(n)$ is the output IN mitigated signal as in (29). Using (37) in (33), the optimal hybrid clipping-blanking threshold can be estimated as presented in Section VI-C.

VI. Results and Discussions

From the ongoing discussion, it can be categorically mentioned that an optimal amplitude threshold at which IN is mitigated to achieve optimal output SNR is desired for PLC-OFDM systems. This can be achieved from converting the conventional Rayleigh amplitude distribution of unmodified OFDM system to a uniform distribution. We tackle this by companding the signal amplitudes to, first, reduce the PAPR before transmission over powerline channels. By computer simulation using MATLAB, we present the procedure and discuss simulation results of the proposed systems as well as the conventional OFDM approach. To achieve these, we generate $N = 4096$ random data first and then modulate them using 16-QAM constellations. The output signals are then scaled to ensure unit output power before passing the output signal through IFFT block to obtain the time-domain signals expressed in (1). When passed through the IFFT block, the resulting amplitudes assumes the Rayleigh distribution described in Section II. To distort the conventional Rayleigh amplitude distribution and reduce the PAPR, we apply the companding transforms discussed in Section III in turns which helps in the mitigation of the IN at the receiver. Then, the output signal is passed through the IN channel with AWGN. To mitigate IN in the received signal, we apply the nonlinear preprocessing to remove the identified IN. Finally, the output result is then used to measure the output SNR for further analyses involving optimization.

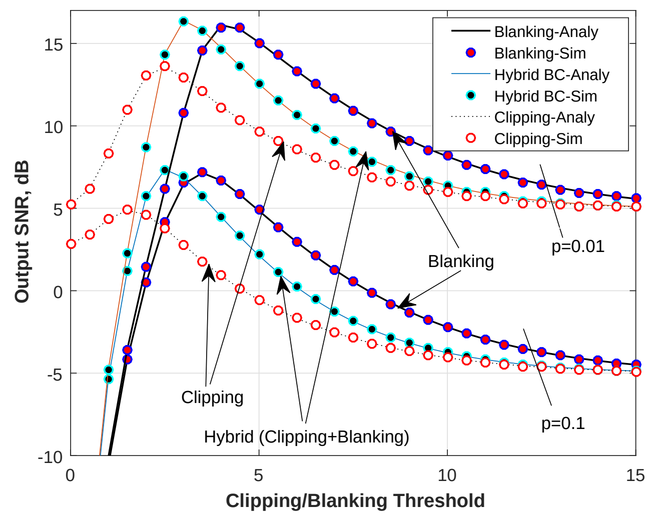

Figure 8. Performance comparisons of received output SNR and clipping/blanking thresholds for simulated system model and analytical model, SNR = 25dB, SINR = -15 dB, $p = 0.01$, $N = 4096$

A. Performance Evaluation of Analytical and Simulated Results in Terms of Output SNR

We verify our system models by comparing the simulated results with the analytical results using the output SNR method described in Section IV-C. To begin with, we compare the output SNR performances with the clipping-blanking thresholds for the three nonlinear IN mitigation techniques in Fig. 8. Starting with the analytical and simulated performances of clipping and blanking schemes which respectively match. The noise incurred in clipping excess amplitudes influences the in-band distortions induced by the clipping scheme thus diminishing the output SNR. This noise may not be present in blanking so that the blanking scheme outperforms the clipping scheme. Their hybrid demonstrates even better performance than either blanking or clipping operating in solitude. Based on these results, the performance optimization to finding the optimal nonlinear mitigating amplitude for which the systems dispenses with maximal SNR shall be continued using the hybrid clipping-blanking technique. Meanwhile, let us emphasize the diminished performance exhibited by increasing the probability of IN occurrence from $p = 0.01$ to $p = 0.1$. Observe also that all simulated results perfectly match with the and analytical results.

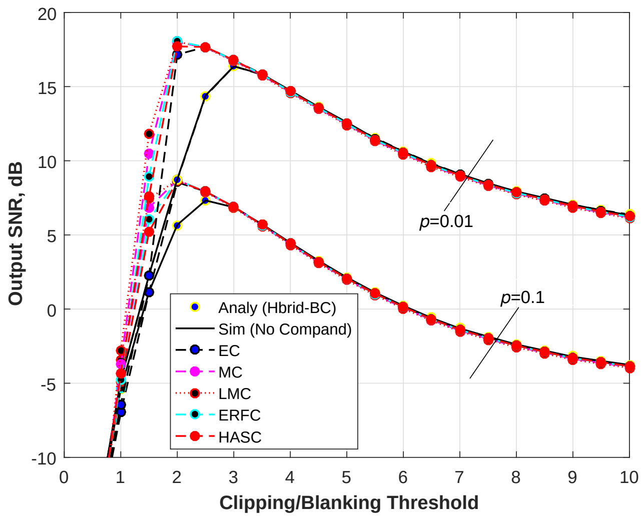

Figure 9. Performance comparisons of received output SNR and clipping/blanking thresholds for different companding models with reference to the analytical model, SNR = 25dB, SINR = -15 dB, $p = 0.01$, $N = 4096$

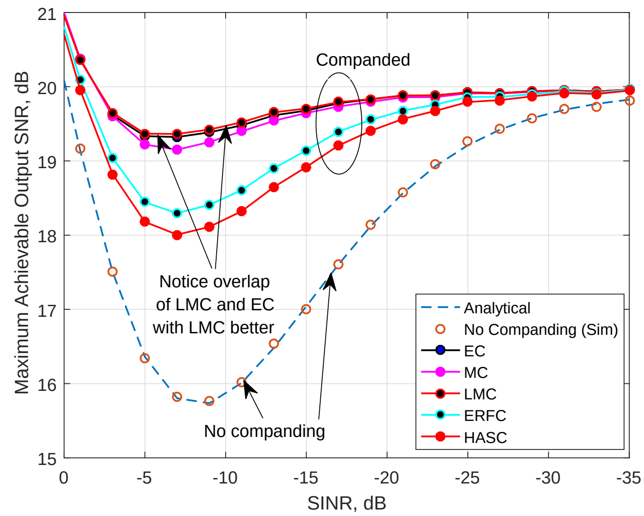

Figure 10. Optimal output SNR of companded OFDM system using hybrid clipping-blanking amplitude thresholds, SNR = 50dB, $p = 0.01$, $N = 4096$

B. Performance Evaluation of the Model System over AWGN Channel with Impulsive Noise

The output SNR measures the output power of the signal against the mixture non-Gaussian noise in the system. In this study, we explore different companding PAPR models that drive an OFDM system towards uniform distribution to reduce the PAPR and increase system performance against IN. Thus, we found useful in the mitigation of IN at the receiver of the PLC system as already shown in Fig. 7. Now, in Fig. 9, we show the output SNR performances of these companding schemes with reference to the analytical and uncompanded model.

Then, recall that in Fig. 3, we presented the different companding transforms in comparisons to one another. Companding PAPR reduction scheme, generally, expands or compresses the amplitudes of OFDM signal or both. All considered companding transforms increase the output SNR beyond that of the unmodified OFDM system as in Fig. 9 as they transform the amplitudes of the system towards uniform distribution and any intrusive amplitude can be clearly observed and removed; increased mitigation of IN translates into increased SNR as the noise power is reduced in this case. However, increasing the probability of IN occurrence, decreases the output SNR although the companded signals consistently dissipate better output SNR than the unmodified system.

In Fig. 9, the PLC system attains the maximum output SNR at a specific clipping/blanking threshold. Now, give the input SINR and p there should exist an optimal SNR provided the PDF point to the optimal amplitude distribution. This optimization problem of the form (33) involves an exhaustive search for the optimal clipping/blanking amplitude.

Recall that the PDF distribution influences the mitigation of IN presence in the PLC system. As an example, when optimization search is performed in terms of the output SNR performances in Fig. 10, the analytical and simulated model for the uncompanded OFDM signals dispensing with the hybrid clipping and blanking perfectly agree. In addition, the performance strengths of the proposed companding techniques on the systems depict that all the companding transforms demonstrate higher output SNRs than the conventional OFDM signals operating without amplitude modification. Recall that in Fig. 5, EC presents output companded signal with all amplitudes centering around the mean amplitude, then followed by the LMC and MC schemes. Consequently, the EC and LMC schemes achieve the most performing SNRs while the ERFC and HASC are the least performing. By varying the percentage of IN, p, we investigate the performance of the system under study for $p = 0.1$ as shown in Fig. 11. However, similar to the trends in Fig. 10, the companded signals outperform the uncompanded signal with LMC and EC achieving the best performances.

In Fig. 9, we found that the output SNR is optimum at only one identifiable amplitude threshold. Also, in Fig. 8, these threshold amplitudes shift closer to unity implying that the most amplitude may exist below the present. Consequently, using the proposed optimization search, we find in Fig. 12 that the amplitudes that exude the optimal output SNRs exist far below the conventional threshold of the unmodified OFDM system. Expressly, the optimal amplitude thresholds for mitigating the impulsive is lower for better performing PAPR schemes with the LMC and EC achieving the least of these amplitude thresholds.

Finally, increasing the percentage of IN occurrence in the system lowers the output SNR. It is worthy to note that the overall output SNR is correspondingly decreased due to the increased noise power. The amplitudes consequently become decreased, however, the LMC and EC PAPR reduction schemes achieve better performances as shown in Fig. 13. This observation holds for the corresponding result in Fig. 11 where the IN probability has been increased from $p = 0.01$ to $p = 0.1$.

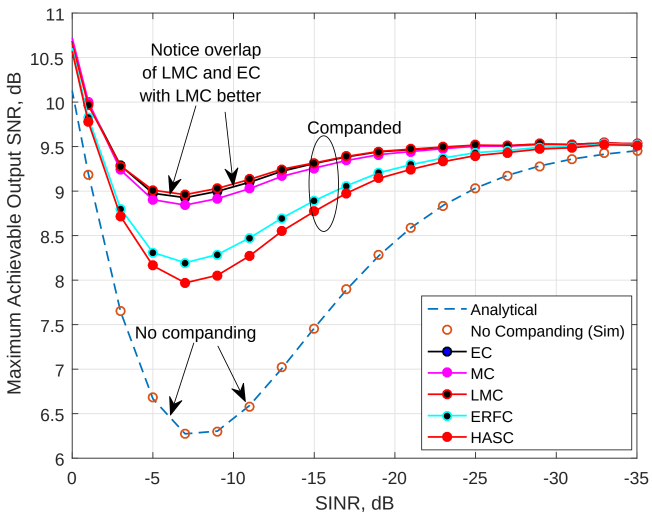

Figure 11. Optimization of the output SNR of the system model under consideration, SNR = 50dB, $p = 0.1$, $N = 4096$

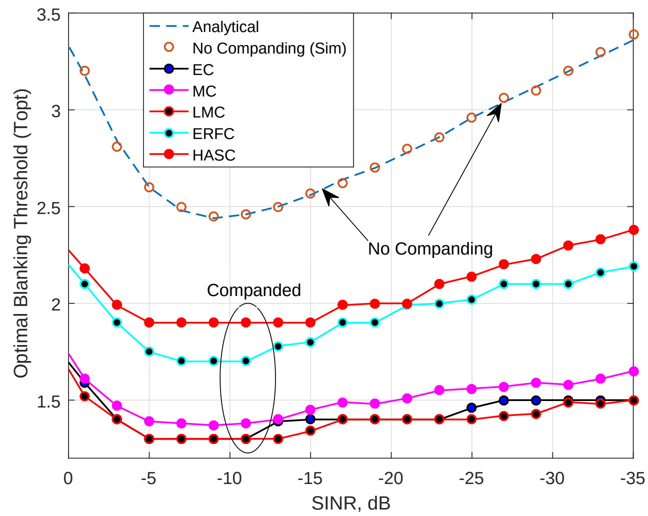

Figure 13. Optimized Clipping-Blanking threshold for mitigating IN in PLC-OFDM system using companding of the system model under consideration; SNR = 50dB, $p = 0.1$, $N = 4096$

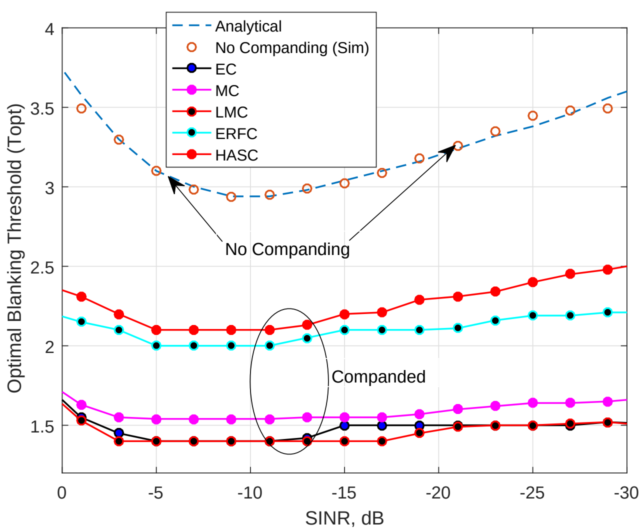

Figure 12. Optimized Clipping-Blanking threshold for mitigating IN in PLC-OFDM system using companding of the system model under consideration; SNR = 50dB, $p = 0.01$, $N = 4096$

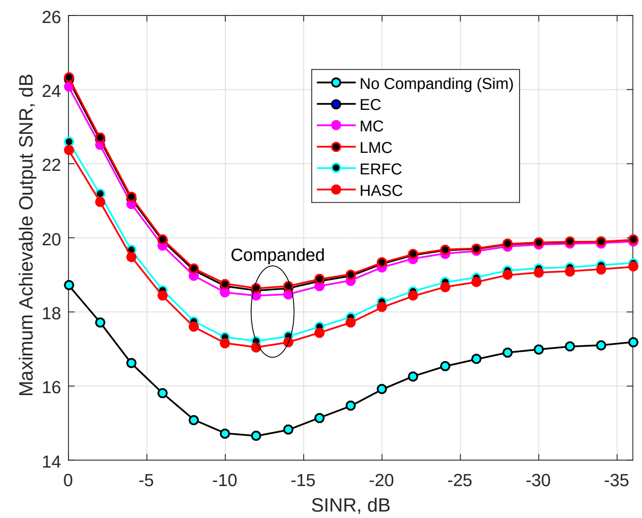

Figure 14. Performance evaluation of optimized IN reduction process for PAPR reduced OFDM signal over PLC fading channel in terms of maximal output SNR, $p = 0.01$

We establish finally that unlike the earlier study that limited the application of companding to MC only for availing the IN mitigation, there are at least two more from the foregoing study that perform far better than the MC scheme. It is possible that among the existing companding schemes, there may exist other models that perform better than the MC scheme in addition to the presented. This is an open research that anyone can further investigate.

C. Performance Evaluation over Multipath Fading Channel with IN and AWGN

Similar to the foregoing discussions, the results of our optimal processes are demonstrated in Fig. 14 for a PLC multipath fading channel described in Section V-A. Multipath fading coefficients attenuate the signal amplitudes and consequently diminish the received output SNR. However, the trend as already observed without fading channel above is consistent. For example, applying companding to the signal before transmission over the channel with multipath achieves 5dB gain in terms of output SNR. Furthermore, comparing Fig. 10 and Fig. 14, it can be observed that the output SNR in the latter is diminished compared to the former.

When the probability of IN occurrence is increased from $p = 0.01$ to $p = 0.1$, the output SNR are shown in Fig. 15. We observe all companded models outperform the uncompanded signals in terms of output SNR. However, the LMC achieves better output SNR than the rest companding models.

In Fig. 16, the amplitude performance corresponding to

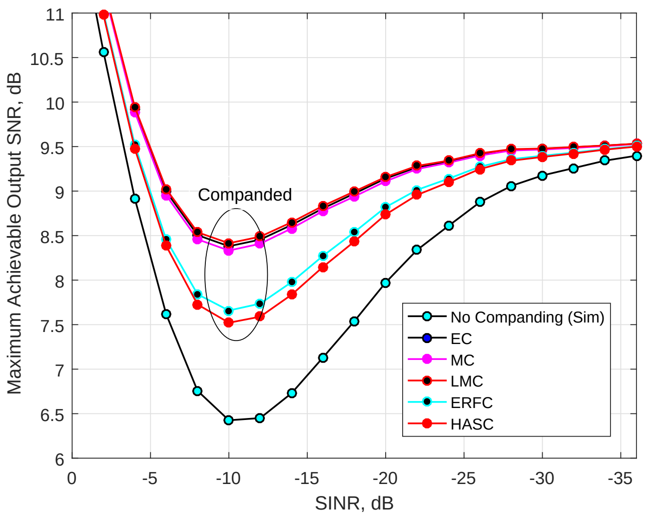

Figure 15. Performance evaluation of optimized IN reduction process for PAPR reduced OFDM signal over PLC fading channel in terms of maximal output SNR, $p = 0.1$

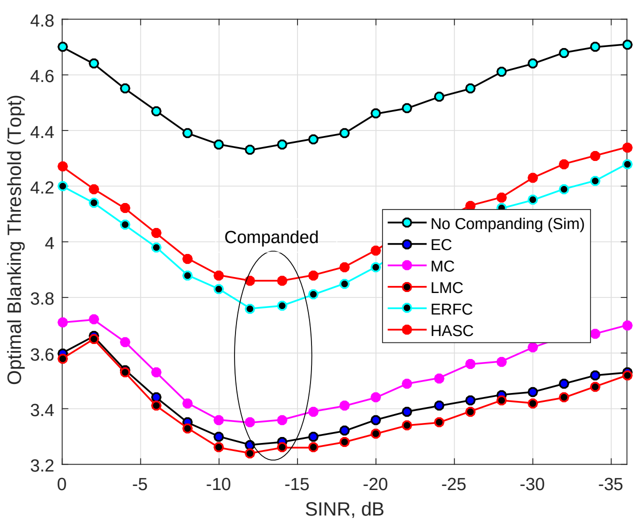

Figure 16. Performance evaluation of optimized IN reduction process for PAPR reduced OFDM signal over PLC fading channel in terms of hybrid clipping-blanking threshold, $p = 0.01$

$p = 0.01$ are shown. We have demonstrated in Section VI-B that clipping-blanking thresholds are inversely proportional to the output SNR as higher amplitude becloud the presence of IN. Thus, we observe in Fig. 16 that companding transforms with better output SNRs in Fig. 14 exude lower amplitude thresholds with LMC being the least. Finally, we consider the hybrid clipping-blanking threshold when the probability of IN occurrence is increased from $p = 0.01$ to $p = 0.1$ as shown in Fig. 17. Increasing the percentage of IN occurrence leads to higher noise power and thus further degrades the output SNR. Consequently, the corresponding optimal mitigation thresholds follow suit with the companding transforms that achieve better output SNR showing lower amplitude thresholds.

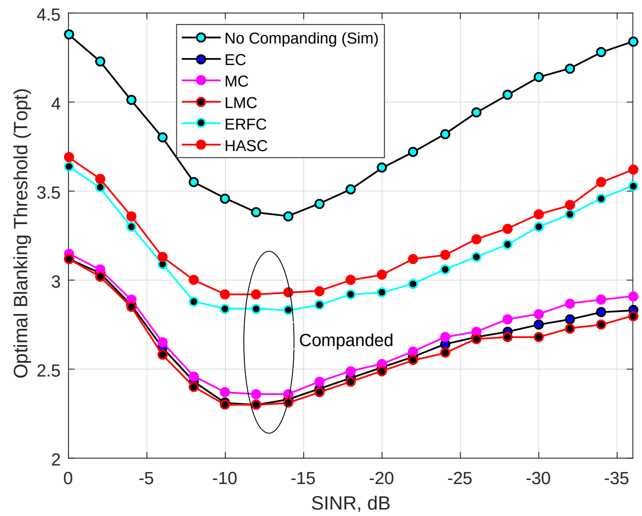

Figure 17. Performance evaluation of optimized IN reduction process for PAPR reduced OFDM signal over PLC fading channel in terms of hybrid clipping-blanking threshold, $p = 0.1$

VII. Conclusion

We have presented a detailed evaluation of different companding PAPR reduction styles to enhance the mitigation of impulsive noise in PLC systems. We are motivated by the fact that any uniformly distributed amplitudes could avail the identification of any intrusive change in the amplitude. Converting the PDF to a uniform distribution presents a better platform for mitigating IN in PLC systems. Using the five different companding transforms namely MC, EC, ERFC, LMC and HASC, we found that the output SNR of nonlinearly mitigated IN in the system can be significantly enhanced. Then, with optimization technique, we explored the optimal amplitude for mitigating the IN to realize the best output SNR performance with hybrid clipping-blanking. Results show that these amplitudes exist far lower than the amplitudes presented by the unmodified amplitude distributed OFDM system. Among the companding schemes, the LMC scheme followed by the EC scheme show the most performing output SNRs due to their closely uniformly distribution of OFDM signal amplitudes around the mean. We found also that the conventional amplitude distribution gives false indication of the presence of IN. Over multipath fading channel, we showed that the attenuation of the signal amplitudes by the channel coefficients degrades the SNR performance compared to simply impulsive channel with AWGN presence.

Acknowledgment

This research has been carried out within the Peer-to-Peer Energy Trading and Sharing - 3M (Multi-times, Multi-scales, Multi-qualities) project funded by EPSRC (EP/N03466X/1).

References

[1] A. Ikpehai, B. Adebisi, and K. M. Rabie, “Broadband PLC for Clustered Advanced Metering Infrastructure (AMI) Architecture,” *Energies*, vol. 9, p. 569, Jul. 2016.

[2] L. Lampe, A. M. Tonello, and T. G. Swart, *Power Line Communications: Principles, Standards and Applications from multimedia to smart grid*. John Wiley & Sons, second ed., 2016.

[3] X. Fang, S. Misra, G. Xue, and D. Yang, “Smart Grid—The New and Improved Power Grid: A Survey,” *IEEE Commun. Surveys Tut.*, vol. 14, no. 4, pp. 944–980, 2012.

[4] M. M. Rahman, C. S. Hong, S. Lee, J. Lee, M. A. Razzaque, and J. H. Kim, “Medium access control for power line communications: an overview of the IEEE 1901 and ITU-T G.hn standards,” *IEEE Commun. Mag.*, vol. 49, pp. 183–191, Jun. 2011.

[5] K. O. Anoh, J. M. Noras, R. A. Abd-Alhameed, S. M. Jones, and K. N. Voudouris, “A new approach for designing orthogonal wavelets for multicarrier applications,” *AEU-Intl. J. Electron. Commun.*, vol. 68, pp. 616–622, Jul. 2014.

[6] K. O. Anoh, T. T. Mapoka, R. A. Abd-Alhameed, O. Ochonogor, and S. M. Jones, “On the Application of Raised-Cosine Wavelets for Multicarrier Systems Design,” *IRECAP*, vol. 4, pp. 143–150, Aug. 2014.

[7] S. V. Zhidkov, “Performance analysis and optimization of OFDM receiver with blanking nonlinearity in impulsive noise environment,” *IEEE Trans. Veh. Technol.*, vol. 55, pp. 234–242, Jan. 2006.

[8] S. V. Zhidkov, “Analysis and comparison of several simple impulsive noise mitigation schemes for OFDM receivers,” *IEEE Trans. Commun.*, vol. 56, pp. 5–9, Jan. 2008.

[9] K. M. Rabie and E. Alsusa, “Efficient SLM based impulsive noise reduction in powerline OFDM communication systems,” in *2013 IEEE GLOBECOM*, pp. 2915–2920, Dec. 2013.

[10] K. M. Rabie, E. Alsusa, A. D. Familua, and L. Cheng, “Constant envelope OFDM transmission over impulsive noise power-line communication channels,” in *2015 IEEE ISPLC*, pp. 13–18, Mar. 2015.

[11] R. Almahainy, N. Ali, S. Al-Araji, and M. Ismail, “Impulsive noise mitigation for μ-law companded OFDM-based communication system,” in *IEEE ICECS*, pp. 264–267, Dec. 2015.

[12] J. Joung, C. K. Ho, K. Adachi, and S. Sun, “A survey on power-amplifier-centric techniques for spectrum-and energy-efficient wireless communications,” *IEEE Commun. Surveys & Tuts.*, vol. 17, pp. 315–333, Firstquarter 2015.

[13] R. Yoshizawa and H. Ochiai, “Effect of Clipping and Filtering with Distortionless PAPR Reduction for OFDM Systems,” in *2015 IEEE 82nd VTC Fall*, pp. 1–5, Sept. 2015.

[14] T. Jiang, C. Li, and C. Ni, “Effect of PAPR Reduction on Spectrum and Energy Efficiencies in Ofdm systems With Class-A HPA Over AWGN Channel,” *IEEE Trans. Broadcast.*, vol. 59, pp. 513–519, Sept. 2013.

[15] J. Joung, C. K. Ho, and S. Sun, “Spectral Efficiency and Energy Efficiency of OFDM Systems: Impact of Power Amplifiers and Countermeasures,” *IEEE J. Sel. Areas in Commun.*, vol. 32, pp. 208–220, Feb. 2014.

[16] K. M. Rabie and E. Alsusa, “On enhancing the performance of the DPTE-based noise cancellation method utilizing the PTS PAPR reduction scheme in PLC systems,” in *18th IEEE ISPLC*, pp. 334–339, Mar. 2014.

[17] K. Anoh, B. Adebisi, and M. Hammoudeh, “A Comparison of ICF and Companding for Impulsive Noise Mitigation in Powerline Communication Systems,” in *Proceedings of ICFNDS ’17, Cambridge-UK*, Jul. 2017.

[18] X. Wang, T. T. Tjhung, and C. S. Ng, “Reduction of peak-to-average power ratio of OFDM system using a companding technique,” *IEEE Trans. Broadcast.*, vol. 45, pp. 303–307, Sept. 1999.

[19] T. Jiang, W. Xiang, P. C. Richardson, D. Qu, and G. Zhu, “On the nonlinear companding transform for reduction in PAPR of MCM signals,” *IEEE Trans. Wireless Commun.*, vol. 6, pp. 2017–2021, Jun. 2007.

[20] T. Jiang, Y. Yang, and Y.-H. Song, “Exponential companding technique for PAPR reduction in OFDM systems,” *IEEE Trans. Broadcast.*, vol. 51, pp. 244–248, Jun. 2005.

[21] J. Hou, J. Ge, D. Zhai, and J. Li, “Peak-to-average power ratio reduction of OFDM signals with nonlinear companding scheme,” *IEEE Trans. Broadcast.*, vol. 56, pp. 258–262, Apr. 2010.

[22] O. Gazi, “A new companding technique for PAPR reduction in OFDM communication systems,” in *3rd ICUMT*, pp. 1–5, Oct. 2011.

[23] Y. Wang, L. h. Wang, J. h. Ge, and B. Ai, “Nonlinear companding transform technique for reducing PAPR of OFDM signals,” *IEEE Trans. Consum. Electron.*, vol. 58, pp. 752–757, Aug. 2012.

[24] P. Banelli, “Non-linear transformations of Gaussians and Gaussian-mixtures with implications on estimation and information theory,” *arXiv preprint arXiv:1111.5950v3*, May 2013.

[25] K. Anoh, C. Tanriover, and B. Adebisi, “On the Optimization of Iterative Clipping and Filtering for PAPR Reduction in OFDM Systems,” *IEEE Access*, vol. 5, pp. 12004–12013, Jun. 2017.

[26] K. Anoh, R. Abd-Alhameed, Y. Dama, S. Jones, P. Pillai, and K. Voudouris, “An investigation of PMEPR of WPT-OFDM and OFDM multicarrier systems,” *J. Commun. Netw.*, vol. 3, pp. 45–52, Jul. 2013.

[27] J. Lin, M. Nassar, and B. L. Evans, “Impulsive Noise Mitigation in Powerline Communications Using Sparse Bayesian Learning,” *IEEE J. Sel. Areas Commun.*, vol. 31, pp. 1172–1183, Mar. 2013.

[28] D.-F. Tseng, Y. S. Han, W. H. Mow, L.-C. Chang, and A. H. Vinck, “Robust clipping for OFDM transmissions over memoryless impulsive noise channels,” *IEEE Commun. Lett.*, vol. 16, pp. 1110–1113, Jul. 2012.

[29] K. M. Rabie, B. Adebisi, and A. Salem, “Improving energy efficiency in dual-hop cooperative PLC relaying systems,” in *ISPLC*, pp. 196–200, Mar. 2016.

[30] K. Al-Mawali, A. Z. Sadik, and Z. M. Hussain, “Joint Time-domain/Frequency-domain Impulsive Noise Reduction in OFDM-based Power Line Communications,” in *2008 Australasian Telecom. Netw. Appl. Conf.*, pp. 138–142, Dec. 2008.

[31] H. Yasui, A. Nakamura, and M. Itami, “Impulsive noise mitigation scheme that combines nulling and time sample replacement,” in *IEEE Intl. Conf. Consum. Electron. (ICCE)*, pp. 337–338, Jan. 2016.

[32] K. M. Rabie and E. Alsusa, “Single-carrier FDMA with blanking/clipping for mitigating impulsive noise over PLC channels,” in *18th IEEE ISPLC*, pp. 340–345, Mar. 2014.

[33] B. Adebisi, K. M. Rabie, A. Ikpehai, C. Soltanpur, and A. Wells, “Vector ofdm transmission over non-gaussian power line communication channels,” *IEEE Syst. J.*, 2017.

[34] M. Rozman, A. Ikpehai, B. Adebisi, and K. M. Rabie, “Channel characterisation of cooperative relaying power line communication systems,” in *Proc. 10th Intl. CSNDSP*, pp. 1–5, Jul. 2016.

[35] M. Zimmermann and K. Dostert, “A multipath model for the powerline channel,” *IEEE Trans. Commun.*, vol. 50, pp. 553–559, Apr. 2002.

[36] B. Adebisi, A. Khalid, Y. Tsado, and B. Honary, “Narrowband PLC channel modelling for smart grid applications,” in *2014 9th Intl. Symposium CSNDSP*, pp. 67–72, Jul. 2014.

[37] S. Güzelgöz, H. B. Celebi, and H. Arsian, “Analysis of a multi-channel receiver: Wireless and plc reception,” in *2010 18th European Signal Process. Conf.*, pp. 1106–1110, Aug. 2010.

Kelvin ANOH (S'11 – M'15) received MSc degree in Data Telecommunications and Networks from the University of Salford, UK in 2010 and then PhD degree in Telecommunications Engineering from the University of Bradford, UK in 2015. He also received BSc (Hons) degree in Industrial Physics from Ebonyi State University, Nigeria in 2006. Since 2016, he is a Research Associate with Manchester Metropolitan University (MMU), UK where he worked on an InnovateUK-EPSRC project that received both the Knowledge Exchange Project and Outstanding knowledge Awards in 2016 respectively. He received the Best Paper award at the University of Cambridge, UK in 2017 during the ICFNDS'17 conference. His research interests are in the areas of signal processing and emerging communication technologies. He is a member of IET and IEEE.

Bamidele ADEBISI (M'06, SM'15) received his Master's degree in advanced mobile communication engineering and Ph.D. in communication systems from Lancaster University, UK, in 2003 and 2009, respectively. Before that, he obtained a Bachelor's degree in electrical engineering from Ahmadu Bello University, Zaria, Nigeria, in 1999. He was a senior research associate in the School of Computing and Communication, Lancaster University between 2005 and 2012. He joined Metropolitan University, Manchester in 2012 where he is currently a Reader in Electrical and Electronic Engineering. He has worked on several commercial and government projects focusing on various aspects of wireline and wireless communications. He is particularly interested in Research and Development of communication technologies for electrical energy monitoring/management, transport, water, critical infrastructures protection, home automation, IoTs and Cyber Physical Systems. He has several publications and a patent in the research area of data communications over power line networks and smart grid. He is a member of IET and a senior member of IEEE.

Khaled M. RABIE (S'12–M'15) received the B.Sc. degree (Hons.) in electrical and electronic engineering from the University of Tripoli, Tripoli, Libya, in 2008, and the M.Sc. and Ph.D. degrees in communication engineering from the University of Manchester, Manchester, U.K., in 2010 and 2015, respectively. He is currently a Post-Doctoral Research Associate with Manchester Metropolitan University (MMU), Manchester. His research interests include signal processing and analysis of power-line and wireless communication networks. He received several awards, nationally and internationally, including the Agilent Technologies' Best M.Sc. Student Award, the Manchester Doctoral College Ph.D. Scholarship, and the MMU Outstanding Knowledge Exchange Project Award of 2016. He was also a recipient of the Best Student Paper Award at the IEEE International Symposium on Power Line Communications and its applications in 2015, TX, USA.

Mohammad HAMMOUDEH is a Senior Lecturer in Computer Networks and Security in the School of Computing, Mathematics and Digital Technology at the Manchester Metropolitan University. He received his Ph.D. in Computer Science from the University of Wolverhampton in 2009, his MSc in Advanced Distributed Systems from the University of Leicester in 2007 and his BSc (Hons) in Computer Communications from the Arts, Sciences & Technology University in Lebanon in 2004. He is the co-founder and member of the FUture Networks and Distributed Systems research Group (FUNDS). He is the founder and head of the MMU IoT Lab.

Haris GAČANIN received his Dipl.-Ing. degree in Electrical Engineering from University of Sarajevo in 2000. In 2005 and 2008, respectively, he received MSc and PhD from Tohoku University in Japan. He was with Tohoku University from April 2008 until May 2010 first as Japan Society for Promotion of Science postdoctoral fellow and then he was appointed as Assistant Professor. Since 2010, he leads Artificial Intelligence Communication Systems department at Nokia Bell Labs. His professional interest is related to application of artificial intelligence to enable autonomous networking of mobile and wireless systems. He has more than 180 scientific publications (journals, conferences and patent applications), invited and tutorial talks. He is senior member of the Institute of Electrical and Electronics Engineers (IEEE) and the Institute of Electronics, Information and Communication Engineering (IEICE).