Enhanced Biased ASK Modulation Performance for SWIPT with AWGN Channel and Dual-Purpose Hardware

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Abstract—A lot of small sensors are deployed in Internet of Things applications. Simultaneous information and power transfer (SWIPT) is a solution to avoid a large amount of batteries, while exchanging information with the sensors. Combining both energy harvesting and information receiving in dual-purpose hardware lowers power consumption while also decreasing sensor size. Biased ASK is introduced as a solution for the SWIPT modulation technique. In particular the design of enhanced biased ASK modulation in case of an additive white Gaussian noise channel is analyzed in this paper. Novel formulas for symbol error rate are derived while improvements of the modulation scheme are proposed based on measurements, taking rectifier nonlinearity into account and resulting in reducing the transmitted power by 14 dB for fixed error rate.

Index Terms—Modulation, Rate-energy tradeoffs, Receiver architecture, SWIPT, WPT.

I. INTRODUCTION

SIMULTANEOUS wireless information and power transfer (SWIPT) is very useful in Internet of Things applications for avoiding a lot of large batteries. In SWIPT literature, wireless power transfer (WPT) researchers focus on rectifier power conversion efficiency [1]–[7], while wireless information transfer (WIT) research focuses on system model designs and energy budget constraints [8]. Neither consider the mutual impact of both subsystems. Some research about the effect of waveforms on both subsystems has been done but is largely limited to theoretical work with typically quite some assumptions on the hardware, without validation by measurements [9], [10].

Interestingly, when relying on amplitude modulation, it is possible to use the rectifier as information receiver, de facto taking care of the conversion from RF to baseband without the need of a power consuming local oscillator. When considering an integrated power and information receiver (or rectifier), it can not be jointly optimized for power and information, as there is a trade-off between both.

The authors of [11] have compared the integrated and separated receiver types for SWIPT systems using simulations. In [12], a PAPR based modulation technique is proposed which is the first modulation technique using the typical WPT multisine waveform, specifically designed for an integrated receiver. However, the feasibility of the proposed scheme was not confirmed by measurements. In [13], the authors propose

another new modulation scheme exploiting the dependency of the output on the amplitude ratio of two-tones at the rectifier input.

In [14], we have introduced a biased ASK modulation technique, enabling information transfer and power transfer in a SWIPT system with dual-purpose hardware. To demonstrate the feasibility, a power optimised rectifier was used as receiver for both WIT and WPT. The main contribution of [14] is the joint modelling of information and power reception performance of an integrated receiver circuit with dual-purpose hardware (HW), and the exploration of the trade-off between both, using the proposed novel biased ASK modulation. Our earlier measurements of modulated multisine signals in [15] have shown that the existence of an output ripple depends on the modulation scheme, symbol rate and lowpass filter characteristics. Hence, in [14] we derived a design guideline for the modulation scheme and rectifier hardware to enable a correct information transfer.

This work extends [14], where we introduced the novel biased ASK modulation technique and parameters F_{ratio} and A_{ratio} , which impact the trade-off between WPT and WIT. In [14], we already demonstrated this trade-off in terms of F_{ratio} using measurements. To analyze also the complete impact of A_{ratio} however, we now introduce a noisy channel, and use measurements to show that generally, low A_{ratio} improves WIT while high A_{ratio} improves WPT. Symbol error rate (SER) is used as a figure of merit for the WIT link. We derive a novel mathematical expression for SER for the biased ASK modulation technique in the presence of AWGN noise, which we compare with simulation and measurement results. Also, we introduce two enhancements to the system setup to further improve performance of the proposed modulation scheme, taking the hardware power dependent nonlinearity into account. In this work we use the same hardware as in [14]. In both works, the novelty can be found in the analysis of the trade-off between WPT and WIT using the proposed novel modulation technique, rather than in hardware, i.e., rectifier, design.

This scheme can be used in systems where either the transmitter controls its waveform and decides to charge the rectifier or transfer information, or the receiver node can vary its lowpass filter's resistor or capacitor gradually, depending on whether its battery has to be charged or it is ready to decode information. This flexibility is lost when a signal with the proposed modulation scheme is received with separate receivers for power and information, by which typically a fixed power splitter is used to divide the signal [11]. However,

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the mathematical expressions for SER remain useful.

The advantage of our proposed biased ASK modulation scheme is receiver simplicity. Amplitude levels can for example be detected more easily and less power consuming compared to performing fourier transformation as needed for the proposed modulations scheme in [13].

The rest of the paper is organized as follows. Section II introduces the considered theoretical model to analyze the dual-purpose HW behavior for different excitations. The theoretical model and assumptions will be introduced, followed by an introduction to the signal parameters and system trade-offs. Then, the impact of a noisy channel on the modulations performance is theoretically studied in Section III, using simulation results for confirmation. Next, experimental results are used to validate the theoretical trade-offs in Section IV, while also proposing two enhancements to the modulation system to improve performance. Finally, conclusions are drawn in Section V.

II. SYSTEM MODEL

A. Rectifier Model

A standard rectifier contains matching networks, a diode and a lowpass filter. For this paper, a simple theoretical rectifier model is used to derive mathematical formulas and do simulations. Later, measurements are performed to confirm the theoretical conclusions. The simple system model is the same as in [14], shown in Fig. 1. This theoretical model considers the input matching to be perfect. Despite that in most WPT-rectifiers the diode is of Schockley type, the model here considers it to be a zero-threshold or switch. The rectifier's

Fig. 1. Simplified rectifier model, with the diode acting as switch and independence between diode and filter [14].

lowpass filter is constructed with a capacitor and resistor, resulting in a cutoff frequency of:

$$f_{\text{cutoff}} = \frac{1}{2\pi\tau} = \frac{1}{2\pi RC}. \quad (1)$$

The individual impacts of the rectifier and capacitor are not considered since the diode's behavior is considered independent of the lowpass filter, neglecting the effect of loading. Using this theoretical model with an ideal envelop detector in order to analyze a modulations technique's performance, is commonly used in information theory [16]. Hence, using this simple model allows for fair comparison of our novel biased ASK's SER mathematical formulas, with those of existing modulation techniques.

B. Signals

This section describes the proposed modulation scheme as introduced in [14]. Its general characteristics are repeated here for clarity.

Since the rectifier hardware will serve a second purpose as envelope detector, amplitude shift keying (ASK) with various orders M is used because phase information would get lost. To ensure that every symbol contains some minimum power, the ASK modulation is biased. Despite that standard ASK modulation starts at zero amplitude, a ratio between minimum and maximum amplitudes (A_{ratio} or A_r) is introduced as parameter and varied between 0 and 1. For fair comparison between modulations, the average symbol power is fixed. Enhancements of this proposed scheme are presented in Section IV to further take the diode's nonlinearity into account. Fig. 2 shows the biased 4-ASK constellation with A_{ratio} of 0, 0.25, 0.5 and 0.75, though A_{ratio} can be any rational number between 0 and 1. The blue dots represent the different symbol amplitudes and the red vertical lines represent the minimum magnitude and hence power that each symbol has. This ensured minimum power increases with increasing A_{ratio} . Fig. 3 shows the noise margin, minimum amplitude and maximum amplitude for biased M-ASK modulation, represented by NM , A_{min} and A_{max} , respectively. Noise margin is defined as half the amplitude difference of neighboring symbols and will be mathematically defined later in (7). The figure shows that for increasing A_{ratio} , A_{min} increases, A_{max} decreases and noise margin decreases.

Fig. 2. Symbol amplitudes (blue dots), indication of minimum symbol power (red vertical lines), A_{max} (full black line) and A_{min} (dotted black line) for biased 4-ASK with different A_{ratio} for fixed average symbol power [14].

Fig. 3. A_{min} (filled), A_{max} (dotted) and noise margin for different A_{ratio} and M for biased M-ASK with different A_{ratio} with fixed average symbol power.

C. Hardware Behavior Trade-off

Considering a dual-purpose receiver with both information receiving and energy harvesting functionalities, it is important to know the hardware behavior dependency on the excitations. In [14], an upper limit is derived for the symbol rate in case of an noiseless channel, which is discussed again briefly in this section for clarity. In Section IV, measurements will confirm the order of magnitude of the symbol rate that determines this behavioral boundary.

The rectifier has a first order lowpass filter at its output, characterized by (1). The filter's time constant τ represents the speed by which the filter charges or discharges. Assume for example that the rectifier is excited by a signal with constant amplitude $A_{\max}$. When this signal is turned off at t_0 , the output will behave as follows:

$$A(t) = A_{\max} \times \exp \frac{-(t - t_0)}{\tau}, \quad t \geq t_0. \quad (2)$$

For WPT, a filter that is slow relative to the excitation is beneficial since the mean output voltage will be high and the voltage ripple will be small. For WIT however, some output ripple is required to be able to obtain the transferred information from the received waveform. The output will be periodically sampled, requiring the filter to follow the signal's envelope at the sampling time instants. This means that even for the worst case symbol sequence, the filter output should cross the necessary decision boundary after each sample period T , which is the inverse of symbol rate f_{symbol} . This is illustrated in Fig. 4. The figure shows the filter's output (green line) and the biased ASK amplitude levels (blue filled horizontal line) and decision boundaries (blue dotted horizontal lines). The periodically spaced sample moments where the symbols (purple dots) are determined by sampling are indicated by vertical orange lines.

Fig. 4. Output of lowpass filter for pulse train excitation (green) with amplitude sampling moments (orange), symbol amplitudes (blue filled) and decision boundaries (blue dotted) for 4-ASK modulation with $A_{\text{ratio}} = 0$ [14].

This WIT requirement is formulated mathematically in [14] for an ASK modulated CW, resulting in the following upperlimit for the symbol rate

$$F_{\text{ratio}} = \frac{f_{\text{symbol}}}{f_{\text{cutoff}}} \leq \frac{-2\pi}{\ln \left(\frac{A_{\min}}{A_{\max}} + \frac{A_{\max} - A_{\min}}{A_{\max}} \times \frac{1}{2(M-1)} \right)}. \quad (3)$$

Equation (3) leads to two conclusions. Firstly, the upper limit for symbol rate decreases for increasing M . This is because discharge speed has to be higher since noise margin is smaller.

Secondly, the upper limit for symbol rate increases with increasing A_{ratio} since minimum and maximum amplitude are closer together, requiring the capacitor to discharge less.

The modulation will cause ripples at the rectifier output, which is used for WIT. For WPT however, these ripples are undesirable. Choosing a large F_{ratio} or A_{ratio} will reduce ripple and power fluctuations but will also limit the received power that is applied to the power dependent matched region of the diode benefiting WPT, while deteriorating WIT. Choosing small A_{ratio} will also lower the input power range at the rectifier input, respecting the power dependent matching of the diode in the rectifier.

This model neglects many practical non-idealities and parasitic effects so the hardware will behave differently than assumed with our simple model. For example, the loading effect of the RC filter on the diode will cause attenuation of the amplitude levels. Since the derived upper limit in (3) only contains amplitude ratios, such a pure attenuation will have no effect on the described upper limit. However, the originally uniformly spread amplitude levels of the ASK modulation will deform as the diode's model is exponentially shaped, as will be shown by measurements in Section IV.

Despite that the rectifier will impose a nonlinear loading onto the low pass filter, which will change the cutoff frequency, we use F_{ratio} as variable in our modulations and measurements. This is because SER and output voltage are highly dependent on the relative symbol rate. The parameter F_{ratio} has been used before in the state of the art [17].

III. INFORMATION TRANSFER SUBSYSTEM

In a SWIPT system, information transfer typically does not require very high throughput. However, analyzing the modulation's performance when the channel is noisy remains required. The theoretical symbol error rate for 2-ASK is already generally well known. In addition, novel SER formulas for higher orders and varying A_{ratio} are derived in this section, when the channel adds white Gaussian noise to the signal (AWGN channel). Next, a comparison is made between the simulated and measured performance of the modulation scheme.

Symbol $S(i, t)$ is defined as

$$S(i, t) = A(i) \times \cos(2\pi f_c t), \quad \text{for } 0 \leq t \leq T_{\text{symb}} \quad (4)$$

with f_c the carrier frequency, T_{symb} symbol duration, and symbol amplitude levels $A(i)$ expressed by

$$A(i) = A_{\max} \left[\frac{i-1}{M-1} (1 - A_{\text{ratio}}) + A_{\text{ratio}} \right], \quad (5)$$

for $1 \leq i \leq M$.

The decision boundary levels $B(i)$ can be expressed by

$$B(i) = A_{\max} \left[\frac{2i-1}{2(M-1)} (1 - A_{\text{ratio}}) + A_{\text{ratio}} \right], \quad (6)$$

for $1 \leq i \leq M-1$.

The noise margin NM is

$$NM = A_{\max} \frac{1 - A_{\text{ratio}}}{2(M-1)}, \quad (7)$$

which decreases for increasing modulation order M and increasing A_{ratio} . The average symbol power P_{symb} can be found using (5)

$$P_{\text{symb}} = \frac{1}{M} \sum_{n=1}^M \frac{A(i)^2}{2} \quad (8)$$

$$= \frac{A_{\text{max}}^2}{2} \frac{2M(A_r^2 + A_r + 1) - A_r^2 - 4A_r - 1}{6(M-1)}.$$

Symbol error rate (SER) p_{ser} is defined by

$$p_{\text{ser}}(i) = \begin{cases} P\{A(i) > B(1)\} & \text{if } i = 1, \\ 1 - P\{B(i-1) < A(i) < B(i)\} & \text{if } 2 \leq i \leq M-1, \\ P\{A(i) < B(M-1)\} & \text{if } i = M. \end{cases} \quad (9)$$

An analytical expression for p_{ser} for a perfect envelope detector is found by assuming Gaussian distributions for $A(i)$, considering an AWGN channel. The distributions and parameters are shown in Fig. 5. This assumption will result in a good approximation for high amplitude levels and cause estimation errors for levels close to zero where the distribution is known to be Rayleigh. The assumption of Gaussian distributions around uniformly spread amplitude levels at the rectifier output is shown to be inaccurate in Section IV-C, where enhancements to the transmitter or receiver are proposed.

Fig. 5. Assumed received biased ASK modulation amplitude distribution with (rescaled) expected amplitudes ($A(i)$), decision boundaries ($B(i)$), noise margin (NM) and filled green areas for SER calculation.

From fig. 5, (9) can be calculated by

$$p_{\text{ser}}(M, A_{\text{ratio}}, \frac{E_s}{N_0}) \approx \frac{2M-2}{M} \times \int_{A(1)+NM}^{\infty} \frac{1}{\sqrt{2\pi}\sigma} \exp\left(-\frac{r-A(1)}{2\sigma^2}\right) dr$$

$$= \frac{2M-2}{M} \times Q\left(\frac{NM}{\sigma^2}\right). \quad (10)$$

Using the relation between energy and power and the noise-formula using noise spectral density N_0 , (10) can be rewritten

as follows

$$p_{\text{ser}}(M, A_{\text{ratio}}, \frac{E_s}{N_0}) \approx \frac{2M-2}{M} \times Q\left(\frac{NM}{\sqrt{P_{\text{symb}}}} \sqrt{\frac{P_{\text{symb}}}{\sigma^2}}\right)$$

$$= \frac{2M-2}{M} \times Q\left(\frac{NM}{\sqrt{P_{\text{symb}}}} \sqrt{\frac{E_s f_{\text{symb}}}{N_0 B_{\text{bp}}}}\right) \quad (11)$$

$$= \frac{2M-2}{M} \times Q\left(\frac{NM}{\sqrt{P_{\text{symb}}}} \sqrt{\frac{f_{\text{symb}}}{B_{\text{bp}}}} \sqrt{\frac{E_s}{N_0}}\right).$$

Equation (11) shows three factors in the Q-function's argument. The first two factors are fixed by design. $\frac{NM}{\sqrt{P_{\text{symb}}}}$ can be calculated using (7) and (8):

$$\frac{NM}{\sqrt{P_{\text{symb}}}} = \frac{\sqrt{3}(1-A_r)}{\sqrt{(M-1)(2M(A_r^2 + A_r + 1) - A_r^2 - 4A_r - 1)}}. \quad (12)$$

The second factor is the ratio between symbol rate and the bandpass filter's bandwidth B_{bp} at the input of the receiver node.

The third factor represents average symbol energy over noise spectral density. It is important to note that in this work symbol energy E_s is defined as total symbol energy, despite the fact that a constant bias is added to the constellation for non-zero A_{ratio} .

Since the second and third factor in the Q-function's argument are independent from M and A_{ratio} , SER will vary according to (12) when comparing different biased ASK modulations. SER will decrease when (12) increases, which is the case for decreasing M and decreasing A_{ratio} .

Using (11), the theoretical SER for $M = 2$ and $A_{\text{ratio}} = 0$ simplifies to

$$p_{\text{ser}}(2, 0, \frac{E_s}{N_0}) = \frac{1}{2} \exp\left(\frac{-1}{2T_{\text{symb}} B_{\text{bp}}} \frac{E_b}{N_0}\right) \quad (13)$$

which matches with the existing theoretical SER-expression for 2-ASK, as expected. Fig. 6 shows the mathematical SER results using (11) and the simulation results obtained using the simplified system model from Fig. 1 for non-zero A_{ratio} , $M = 2$ and $\frac{B_{\text{bp}}}{f_{\text{symb}}} = 2$.

Equation (11) represents the symbol error rate. This is the same as bit error rate (BER) for 2-ASK. Finding a mathematical expression for bit error rate is often considered to be too complex for higher order modulations. For higher order ASK modulations, symbol error rate can be used as upperlimit for the bit error rate. Grey coding is assumed to minimize the amount of bit errors per symbol error.

IV. EXPERIMENTAL ANALYSIS

Measurements are used to demonstrate the performance of the proposed novel modulation scheme with real rectifier hardware, with and without the presence of an AWGN channel. First, the setup is introduced. Next, the BER and output power are measured. Finally, the AWGN channel is introduced and

Fig. 6. SER simulation and mathematical results for biased 2-ASK modulated CW with A_{ratio} of 0, 0.25, 0.5 and 0.75, including theoretical SER for $M = 2$ and $A_{\text{ratio}} = 0$, with $B_{\text{bp}}/f_{\text{symp}} = 2$.

two enhancements to the modulation scheme are proposed to further improve WIT performance.

A. Measurement Setup

The measurement setup, depicted in Fig. 7, is the one used in [14]. The Vector Signal Transceiver (VST) (NI PXIe-5645R) generates the M-ASK modulated signals and its output is connected to the rectifier. The rectifier has a single-diode circuit topology with an RC-lowpass filter at its output with cutoff frequency 192.92 kHz. The rectifier specifications and design can be found in detail in [18]. The rectifier's baseband output is connected to the VST's baseband receiver which samples the signal to analyze the output's transient behavior, as was enabled by [19]. Since the VST's input port is 50 Ω terminated, it is connected in series with the rectifier's load resistor to limit the impact on the rectifier behavior. This creates a voltage divider which is compensated for in software. The measurements are performed on 20 groups of 10000 random symbols. Carrier frequency is set to 2.45 GHz, transmit power to -15 dBm and symbol rate is 8 kHz, unless specified otherwise.

B. Biased ASK Performance Measurements

1) *Without Channel Noise:* Fig. 8 shows the measurement results for an unbiased ASK modulated CW with varying M and F_{ratio} without channel noise and hence no bandpass filter. These measurements show that for increasing M , WIT performance starts to decrease at smaller F_{ratio} . This corresponds to the theory, described by (3). System nonlinearities and hardware noise cause some deviations, which have more impact on higher order modulations because of their smaller noise margin, calculated with (7). Compensating for these hardware non-idealities is discussed later. The BER measurement results for a 2-ASK modulated CW with varying A_{ratio} are shown in Fig. 9(a). A_{ratio} is varied between 0, 0.25 and 0.5. WIT performance starts to decrease rapidly for F_{ratio} of 9 and 14 for A_{ratio} of 0 and 0.25 respectively, which corresponds to what was found with (3), based on the

Fig. 7. System setup with the VST as RF signal generator and baseband signal receiver, the dual-purpose rectifier, and the voltage divider to connect the rectifier's output to the VST.

Fig. 8. 1-BER measurement results for 2-, 4- and 8-ASK modulated CW with A_{ratio} of zero [14].

simple WIT model in Fig. 1, as shown in [14]. Fig. 9(b) shows that WPT efficiency improves with higher A_{ratio} and F_{ratio} . Improved WPT efficiency with higher F_{ratio} is in line with the results for multisine signals in [15]. By comparing Fig. 9(a) and (b), we see that there is a trade-off between power and information transfer efficiency, depending on F_{ratio} . The downside of increasing A_{ratio} is the decreasing noise margin as shown in Fig. 3. This causes a drop in WIT performance for noisy environments as is discussed in the next section.

The measurement results show a realizable throughput close to 10×192.92 kbps at $\text{BER} = 10^{-3}$ for an unbiased 2-ASK modulated CW excitation on the used rectifier. Hence, rectifier designs with large f_{cutoff} enable larger WIT throughputs but require larger bandwidths for efficient WPT.

2) *With Channel Noise:* In this section we evaluate the derived error rate expressions in Section III by introducing an AWGN channel with variable noise power density to the setup. The transmitter simulates the RF channel output by adding lowpass filtered white Gaussian noise to the signal before upconverting and sending it to the rectifier by wire.

Fig. 10 shows BER measurements for 2-ASK with various

Fig. 9. Measurement results of 1-BER (a) and V_{dc} (b) for 2-ASK modulated CW with variable A_{ratio} [14].

A_{ratio} . These measurements approximate the theoretical result from Section III, certainly for high A_{ratio} . The results also confirm that modulation schemes with larger A_{ratio} are more sensitive to noise.

Fig. 10. BER measurement results (full) and theoretical BER (dotted) for 2-ASK modulated CW with A_{ratio} of 0, 0.25, 0.5 and 0.75 while varying SNR, with $B_{bp}/f_{symb} = 2$.

BER en DC output voltage measurements are shown in Fig. 11 where A_{ratio} is varied while changing symbol rate. Assuming $E_b/N_0 = 100$ to represent the noiseless case, BER increases at smaller symbol rates for modulations with small A_{ratio} as expected from (3). When introducing noise ($E_b/N_0 = 18$), this BER increase happens more gradually but starts at smaller symbol rate, rendering the transition between rectifier-mode and envelope detector-mode less distinct. The BER measurements show that for 2-ASK, $A_{ratio} = 0.5$ (black) provides better WIT performance at higher symbol rates for the noiseless scenario compared to $A_{ratio} = 0.25$ (green), however, introducing noise will lead to a worse performance. Hence WIT performance both depends on the modulation's sensitivity to noise and the impact of the lowpass filter. The mathematical upperlimit from (11) remains a valid upperlimit since noise will never improve performances for higher F_{ratio} . Measuring output voltage shows that introducing high power noise improves power transfer, as expected from [20] since the waveform becomes more chaotic.

Fig. 11. 1-BER and output voltage measurement results for 2-ASK modulated CW with different A_{ratio} , while varying SNR between 18 dB and 100 dB, with $B_{bp}/f_{symb} = 8$.

C. Modulation Scheme Enhancements

In communications theory, the amplitude distribution at the output of a noncoherent receiver is assumed to be the same as the input signal for modulations like ASK. This is also one of the assumptions for deriving (11). However, the diode is nonlinear, which distorts this distribution. In our measurement setup, this caused the 8-ASK modulation to be non-detectable even for the noiseless case. The proposed adaptations in this section are proven to solve this issue. Fig. 12

Fig. 12. Measured output symbol distribution (black dots) and calculated decision boundaries (blue crosses) for 8-ASK modulated CW with A_{ratio} of 0.

shows the received constellation in black dots for different total transmit power levels. Since transmit power varies, output voltage levels will also vary. Hence, the receiver decision boundaries can not be fixed for applications where the receiver sensor, transmitter, or even environment is changing a lot. The expected theoretical decision boundaries can be found after applying the k-means algorithm with $k = M$ on the received constellation in Fig. 12 to determine the M symbols location centers and using (5) and (6)

$$B_{output}(i) = A_{avg.} \frac{A_{max} \left[\frac{2i-1}{2(M-1)}(1 - A_{ratio}) + A_{ratio} \right]}{\frac{1}{M} \sum_{n=1}^M A(i)} \quad (14)$$

$$= A_{avg.} \frac{\left[\frac{2i-1}{2(M-1)}(1 - A_{ratio}) + A_{ratio} \right]}{\frac{(r+1)}{2}},$$

with A_{avg} , the average received symbol power. The decision boundaries are shown in blue marks in Fig. 12. It can be seen that the diode has not only scaled the received symbols but also deformed the signal in a way that the output symbols are not uniformly spaced any more. Hence, using the uniformly distributed decision boundaries will result in a high BER.

Two enhancements to the system model are introduced next to improve BER performance. While still working with a decision making process that is based on euclidean distance, precompensation or postcompensation is employed to cope with the diode's impact on the symbol distribution. Postcompensation requires changes in the receiver while precompensation takes place in the transmitter. Both enhancements take the overall distortion into account, also compensating for transmitter or other receiver nonlinearities and not only the diode.

With postcompensation, the expected constellation in the receiver is changed from the uniformly spread distribution from Fig. 2 to the observed deformed distribution, measured in Fig. 13. A calibration phase is required to obtain these expected levels. The received symbol locations are found using the k-means algorithm on the sampled received signal, with $k = M$. Also, while the AWGN channel introduces Gaussian distributed noise around each amplitude level in the channel as assumed in Section III, the diode will also deform these distributions. This makes the euclidean distance based decision making non-ideal but this is neglected further since performance already improves a lot.

Precompensation takes place in the transmitter. The symbol amplitude levels are designed such that the rectifier output amplitude levels are uniformly distributed as shown in Fig. 2. This is done by varying the transmitted symbol distribution while monitoring the received distribution.

Since the diode has an exponential characteristic, Fig. 13 shows that for postcompensation, the received symbol distribution is compressed towards lower magnitude levels since high amplitudes are amplified more than small amplitudes. Hence, for precompensation the symbol magnitude levels should be increased, compressing the distribution towards higher amplitude levels.

Fig. 13. Measured symbol magnitude distribution with non, pre- or postcompensation for different orders of ASK modulation and A_{ratio} of 0, with $B_{\text{bp}}/f_{\text{sybm}} = 12.5$ at -15 dBm transmitted power.

tions for the proposed integrated receiver. For 4-ASK (green) and BER of 10^{-3} , postcompensation and precompensation improve performance by 12 and 14 dB, respectively. For 8-ASK (pink), pre- and postcompensation are also shown to improve the performance while the non-compensated 8-ASK modulation is not shown since BER was too high (above 5 %) for a lock by the receiver.

Fig. 14. BER measurement results for 2-, 4- and 8-ASK modulated CW with A_{ratio} of 0 while varying symbol rate and comparing pre- and postcompensation.

To employ one of these two calibrations, a calibration phase is needed. The diode's transfer function is power dependent so its impact is too. Hence, a calibration is needed for each power level that is used in the system setup. However, for small power variations, the calibration remains useful as shown in Fig. 15 where BER is measured while varying total received power and using the calibration obtained at -15 dBm. Individual SNR levels are chosen to give comparable BER for each modulation and compensation type at the reference power level of -15 dBm. The variation of BER when varying power in Fig. 15 is not due to noise but to the distortion of the symbol distribution because of the power dependent rectifier nonlinearity. For ideal calibration, BER decreases for increasing SNR as was shown by measurements in Fig. 14. The measurements demonstrate the need for extra calibrations when the used power level range becomes large. A combination of calibrations at selected power levels and interpolation in between is also possible. In case of a time-variant scenario where the power level range is expected to be large, the rectifier-node could perform adaptive postcompensation. The rectifier node could easily measure its mean output voltage and perform adaptive postcompensation, choosing the right calibration coefficients from a pre-installed codebook. Alternatively, a live k-means search algorithm could be performed to calculate the correct calibration during runtime. Though the benefit of compensation is demonstrated for 4- and 8-ASK with A_{ratio} of 0, it is also beneficial for all orders when A_{ratio} differs from zero.

V. CONCLUSION

The novel (enhanced) biased ASK modulation technique is analyzed further, to enable both information receiving and energy harvesting using a power optimized rectifier, avoiding the use of a power consuming local oscillator. The hardware's

Fig. 14 shows the performance of the described compensa-

Fig. 15. BER measurement results for 4- and 8-ASK modulated CW with A_{ratio} of 0 while varying total transmit power and using the calibration obtained at -15 dBm.

behavior is dependent on the power level, noise level, amplitude ratio and ratio $f_{\text{symbol}}/f_{\text{cutoff}}$. The novel mathematical expression for symbol error rate is derived in case of an AWGN channel. The conclusions based on this formula are confirmed by simulations and measurement results. We showed that generally, low F_{ratio} and A_{ratio} improves WIT while high F_{ratio} and A_{ratio} improves WPT.

Enhancements to the biased ASK modulation technique are proposed to cope with circuit nonlinearity and BER performance is studied in the presence of an AWGN channel. Both precompensation and postcompensation of the hardware's nonlinearity lead to an improvement of information transfer but requires a calibration phase for each employed power level. Based on measurements, these compensations can reduce transmitted power by 14 dB for fixed SER.

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