Multitone FSK Modulation for SWIPT

Steven Claessens, *Ning Pan*, *Dominique Schreurs*, and *Sofie Pollin*

***Abstract*—SWIPT is gaining interest as it enables receivers with smaller batteries or none at all. However, the information and power transfer subsystems influence each other's performance, resulting in a trade-off. SWIPT receivers should be low power and compact, ideally resulting in an energy-efficient design where hardware is maximally shared between information and power reception blocks. We propose a novel FSK modulation scheme that uses waveforms which are known to improve wireless power transfer efficiency and can be decoded by relying on the non-linearity of the energy harvesting hardware. This hardware sharing saves area and avoids using a local oscillator. We analyze two novel multitone FSK schemes, while presenting measurement results proving the feasibility of the proposed schemes. Information transfer is based on the relation between input frequency spacings and rectifier output intermodulation product frequencies. We show that contrary to the existing amplitude variation based modulation schemes for SWIPT, the symbol rate for uniform multitone FSK in the noiseless case is not limited by the rectifier low pass filter. For our proposed non-uniform multitone FSK scheme, symbol rate is limited but spectral efficiency is much higher. We show by measurements that information transfer of at least $18\ Mbps$ using the rectifier as receiver is feasible, which is a large increase compared to other solutions. A proposed optimization increases feasible symbol rate for the non-uniform FSK scheme by at least a factor four.**

***Index Terms*—Energy harvesting, Frequency shift keying (FSK), Receiver architecture, Simultaneous wireless information and power transfer (SWIPT), Wireless power transfer (WPT).**

I. Introduction

SIMULTANEOUS wireless information and power transfer (SWIPT) is the combination of wireless information transfer (WIT) and wireless power transfer (WPT). SWIPT is a promising solution to lower the dependency on batteries of sensors, for example in an Internet-of-Things (IoT) network. WIT has been studied extensively. Far field WPT research, however, has recently gained some attention with the research community focusing on the different aspects of a WPT system, trying to improve RF-to-DC power conversion efficiency (PCE) as much as possible. The waveform is also shown to strongly impact PCE. The works in [1]–[7] showed that high PAPR waveforms improve PCE at low input power levels, for far field, radiative WPT. Multitone signals are popular WPT waveforms since their peak-to-average power ratio (PAPR) can easily be controlled by varying the amount of tones. Recently, multisine signals were also adopted for near-field WPT systems [8], which typically uses single tone signals, modulated with amplitude shift keying (ASK), phase shift keying (PSK), frequency shift keying (FSK) or pulse width modulation (PWM) [9], [10]. When adopting WPT efficiency increasing multitone signals, however, these single tone modulation techniques need adaptation. Hence, this works propose a multitone version of classical FSK, where the information is in the distance between tones, enabling power optimized multisine signals for WPT that can be received without LO. The technique is demonstrated using far field WPT hardware, but is also applicable in near field applications.

Despite the research on both topics, combining both the WIT and WPT subsystems remains a challenging topic, especially when the waveform is shared to simultaneously carry both power and information. The receiver can have separated paths for power and information, or integrated paths [11]. Since RF power harvesting is limited due to regulatory constraints and PCE limits, the sensor should be low power. Hence, modulation techniques that enable SWIPT receivers with limited hardware and an alternative for the power consuming local oscillator are very desirable since the considerable decrease in power consumption of the SWIPT receivers [11]. In addition, the SWIPT waveform should adopt the WPT waveform optimizations, so it should have a high PAPR character. Also, it has been shown that there exists a strong mutual impact between both subsystems, which should be taken into account when designing SWIPT modulation techniques [12]. For example, measurement based analysis in [12] showed that for a SWIPT system, WIT performance degrades when high PAPR waveforms are modulated with for example phase shift keying or quadrature amplitude shift keying. This is because of transmitter non-linearity, which causes distortion, especially when transmitting high PAPR waveforms at high power levels. Hence, concerning modulation and waveforms, there is a lot of room for improvement, when comparing to a classical ASK modulated carrier for SWIPT.

Recent work has advanced the practical study of unconventional, optimized SWIPT modulation techniques that can be received without local oscillator. Firstly, the work in [13] introduced a modulation technique that leverages the PCE improvement of multitone signals and its relation between number of tones and PAPR. The information is encoded in the number of tones, while PAPR is measured at the rectifier output. Next, [14], [15] optimized SWIPT waveforms by proposing a biased amplitude shift keying (ASK) modulation scheme where each symbol is ensured to carry some minimum power, trading of WIT for WPT. Finally, the works in [16], [17] proposed encoding the information in the amplitude or phase ratio between two tones of a multitone signal. However, these modulation techniques either impact the rectifier output amplitude in some way or limit the number of input tones,

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risking high amplitude variations.

As a solution, we investigated in [18] the possibility of encoding information in the two tone signal's frequency spacing in order to limit this impact. In addition, output power is shown to increase and be more stable for a high ratio ($F_{\text{ratio}} = \Delta f / f_{\text{cut}}$) between symbol rate and the rectifier's cut-off frequency [12], [19], so a modulation technique that is still decodable by the rectifier at high F_{ratio} is desirable. Frequency shift keying (FSK) is a classical technique for WIT. However, it limits the signal to a single tone, which doesn't optimize PCE. In [18], we extended the FSK technique to two tones and encode the information not in the tones' frequency but in the frequency spacing between the tones. The reason is that the intermodulation products's frequencies at the rectifier output are related to the frequency spacings of the input signal [20]. The information can be decoded without a local oscillator by simply detecting the second intermodulation product's frequency at the rectifier output using the system depicted in Fig. 1. This two tone FSK technique allows for the transmission of information while optimizing PCE by using a multitone signal. Our two-tone FSK technique was shown to increase this decodable F_{ratio} limit, compared to the biased ASK technique. Yet, the proposed novel modulation technique in [18] was based on, and limited to, only two tones, posing a limit on PCE since PAPR is limited to four.

In order to alleviate this limitation, this work is an extension of our two-tone FSK technique in [18], providing two novel, practical solutions to increase the number of tones, while still encoding information in the signal's frequency spacing, using multitone FSK for SWIPT.

The first proposed multitone FSK technique in this work varies the frequency spacing between all N tones at the same time. This is why we call it uniform multitone FSK. It is a straightforward increase of the number of tones, starting from our two-tone FSK modulation technique proposed in [18], allowing for higher PAPR waveforms.

The second proposed technique is called non-uniform FSK, since the frequency spacings between different tones are not the same. This lowers the amount of tones in a certain bandwidth, lowering WPT, but increases spectral efficiency as we will show later in this work. Hence, both PAPR is increased and spectral efficiency is increased compared to the two-tone FSK modulation technique proposed in [18].

Both techniques encode information in the frequency spacing between tones, resulting in sets of multisine symbols with constant PAPR, lowering impact of WIT on WPT efficiency.

This novel and the aforementioned LO-less SWIPT modulation techniques rely on the diode for downconversion, rendering the channel highly non-linear. For these types of channel, the classical Shannon capacity bound is inaccurate [21]. Nevertheless, the relation between system parameters and optimal throughput is analyzed in detail in this work. Because of the channel non-linearity and the fact that no power splitting or time switching is applied, existing optimizations for SWIPT systems, like resource allocation [22], power splitting and time switching optimizations [23] and optimizations of modulation techniques that require a LO at the receiver [24], need revision and new optimizations are likely to arise in the near future to further improve performance of such integrated LO-less SWIPT systems. For example, multiple modulation techniques could be combined in multi-mode SWIPT, depending on the environmental and operating conditions like the power level [25].

The rest of the paper is organized as follows. Section II describes the receiver including the model for simulations and hardware setup for measurements. In Section III, we present both the uniform and non-uniform version of multitone FSK as novel modulation techniques for SWIPT. Simulations and experimental analysis of both schemes are presented in Section IV and Section V. Next, we propose an optimization to improve performance in Section VI, followed by the conclusion in section VII.

II. Receiver model

The system model is shown in Fig. 1. This topology consists of an input matching network, and a diode followed by an RC-low-pass filter (LP-filter), a fourier transformation (FFT) and a bin-detector block. A pass-band filter is placed in front of the rectifier to filter out unnecessary noise and interference. In practical realizations, a passive filter bank could be implemented to realize the FFT operation and further decrease power consumption. The FFT and bin detector are used to demodulate the signals after the integrated rectifier. The bin detector compares for each symbol the energy levels of the baseband bins to find the strongest intermodulation products. Optimizing the hardware and system parameters for WPT limits WIT since ripple and information content will decrease and the rectifier will attenuate the baseband bins non-uniformly. The considered transmitted signal can be described

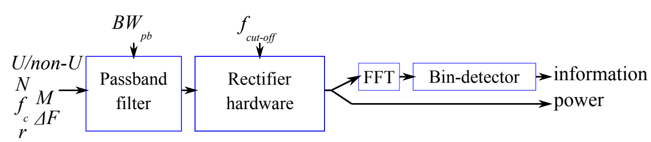

Fig. 1. Receiver system diagram for receiving the proposed multi-tone FSK modulation scheme using FFT.

by

$$S_{in}(t) = \sum_{i=1}^{N} \cos\left(2\pi t \times f_{tone}(i)\right), \qquad (1)$$

with N the number of tones and $f_{tone}(i)$ will be defined in (2) and (3), for the uniform and non-uniform scheme, respectively. The signal $S(t)$ is a sequence of symbols. While the duration of each symbol is constant, they each have a unique set of $f_{tone}(i)$. The diode is simulated by a basic switch model for simulations and the low-pass filter is modeled by transfer function H in (18). Hence, the simulated output signal could be described by $S_{out}(t) = H\left(\max\left(0, S_{in}(t)\right)\right)$.

A. Simulation Model

In this work's simulations, a simple model, which is commonly used in communication theory textbooks, is used to represent the rectifier envelope detector hardware. The diode is considered to behave as a zero-threshold switch, which stops negative signals but passes positive signals. The diode is followed by an RC-low-pass filter, as shown in Fig. 2. This

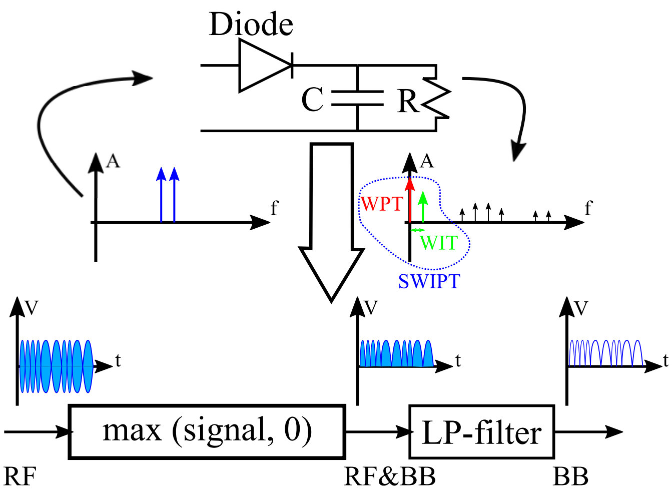

Fig. 2. Circuit model (top) and system model (bottom) for the rectifier hardware, demonstrating the coexistence of a power and IM_2 tone, indicating radio frequency (RF) and baseband (BB) signals.

model was also used in [14], [15] where the simulation results based on this model was shown to approximate WIT measurement results, when considering a (biased) ASK modulation for SWIPT applications. In this model, the low-pass filter and diode are assumed to be independent, meaning that loading effects are ignored.

The rectifier's low-pass filter consists of a parallel resistor and a capacitor. The rectifier's cut-off frequency is f_{cut}.

B. Experimental setup

The measurement results presented in this section verify the relation between input symbol and output second order intermodulation (IM_2) tones and its ability to use them as described in our proposed multitone FSK scheme. The measurements are performed using a vector signal transceiver (VST) (NI PXIe-5645R) for transmitting the 2.45 GHz RF signals. Its RF output is connected to a rectifier which consists of one diode and an RC-low-pass filter. Rectifier specifications are described in [26]. The rectifier's output is connected to the VST's baseband input port using a small voltage divider circuit to avoid loading effects. This setup is shown in Fig. 3. Symbol error rate (SER) and output voltage are measured while varying the modulation parameters in order to study the trade-off between both.

III. Signal Model

We propose two multitone FSK SWIPT modulation schemes: uniform multitone FSK and non-uniform multitone FSK. Both schemes are based on the relation between a multitone's frequency spacings and the rectifier output's intermodulation product's frequencies. Only the frequency spacings are varied, assigning constant and uniform amplitude and phase to all tones in both schemes. The two schemes are discussed in the following sections.

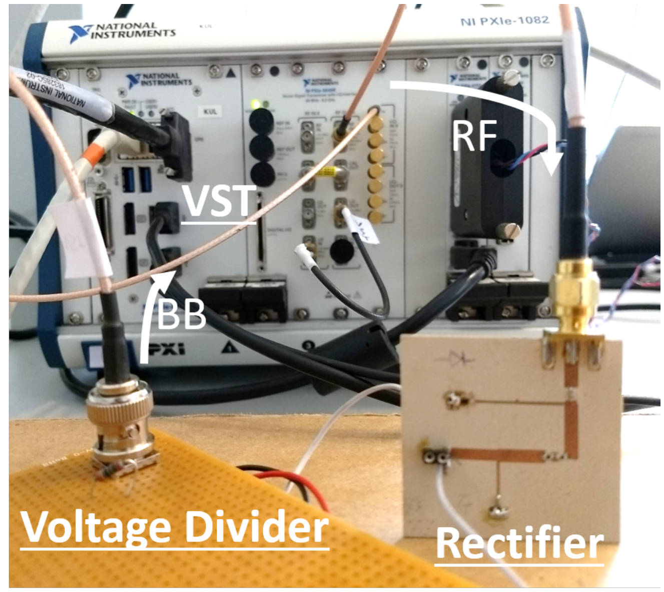

Fig. 3. Measurement setup with VST, rectifier and voltage divider.

A. Uniform Multitone FSK

The idea of uniform multitone FSK is that when transmitting a multitone signal with N tones and uniform frequency spacing Δf between tones, the rectifier output will contain a strong frequency component (IM_2) at frequency Δf, as shown in Fig. 2. Hence, when varying the frequency spacing of the input signal, IM_2 will also shift frequency, which is detectable by using an FFT. This described scheme is called uniform multitone FSK since the frequency spacing between all neighboring tones is the same (multiple of) Δf.

We have demonstrated the feasibility of the uniform 2-tone FSK scheme in [18]. This scheme can be extended to more tones, which is desirable for WPT since input PAPR will increase. For WIT, however, there will not only be one IM_2 tone present at the rectifier output. A lot of different intermodulation products at frequencies, which are multiples of the frequency spacing will occur since not only neighboring tones intermodulate. This poses a challenge in detecting IM_2 with the detector described in Fig. 1. The feasibility of such a uniform multitone scheme depends on the assumption that, for example for the first symbol, IM_2 at $f = \Delta f$ is strongest in amplitude, which is expected since its amount of intermodulating tone pairs in the input signal is the highest.

The uniform multitone FSK frequencies are mathematically determined by

$$\begin{aligned} f_{tone}(n, m, M, r, \Delta f) &= \\ f_c + \left(\frac{1+mr}{2}\right)\Delta f \times (2n - N - 1)&, \\ \text{for } 1 \leq n \leq N&, \\ \text{for } 0 \leq m \leq M-1&, \end{aligned} \tag{2}$$

with r determining the change in frequency spacing between different symbols, n the tone-index starting on the left in the spectrum, N the total amount of tones, M the modulation order and m the symbol-index. This proposed uniform modulation scheme is shown in Fig. 4 for a three tone signal and $M = 4$. The figure clearly shows how different symbols are constructed by only changing the frequency spacings in the same way.

In this uniform scheme we can increase the amount of tones while still keeping the frequency spacing the same between all tones. This restriction can be lifted to increase spectral efficiency for WIT, while still using multiple tones for WPT. We call this non-uniform multitone FSK, which is introduced next.

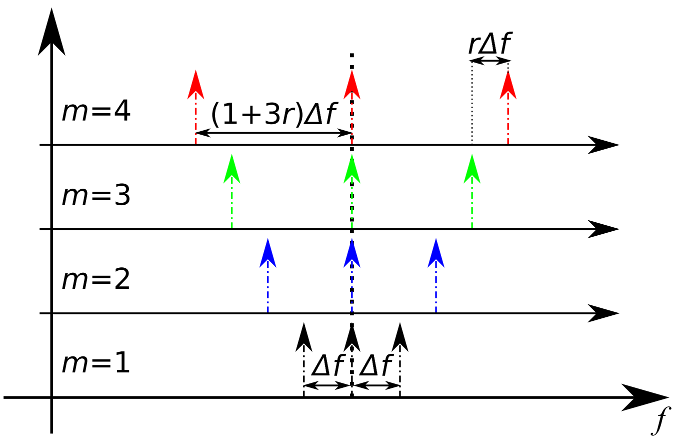

Fig. 4. Signal model illustrating the symbol construction for the proposed novel uniform FSK scheme, illustrated for 3 tones and M=4.

B. Non-Uniform Multitone FSK

In the non-uniform multitone FSK scheme, frequency spacings between neighboring tones are not the same. WPT is expected to be lower since the bandwidth is not optimally, or maximally, filled with tones, but WIT is expected to improve since the amount of bits per symbol increases, increasing spectral efficiency as shown later. For the uniform multitone FSK scheme, for example for the first symbol, $N - 1$ intermodulating tone pairs contribute to IM_2 at $f = \Delta f$ while $N - 2$ intermodulating tone pairs contribute to the intermodulation product at twice that frequency and so forth. In this non-uniform multitone FSK scheme, this is not the case since different frequency spacings are used within one symbol. This proposed non-uniform modulation scheme is illustrated in Fig. 5 for a three tone signal and $M = 4$.

The non-uniform multitone FSK scheme is constructed by choosing tones with different frequency spacings. The amount of frequency spacings considered for N tones is $N(N-1)/2$. These are chosen to assure that the resulting intermodulation products are all located at different frequencies in order to easily detect the original signal based on the rectifier output. Instead of a mathematical solution, we find the desired tone distribution in terms of N by exhaustive search. The frequency spacings for non-uniform 3, 4 and 5-tone FSK are presented, from left to right, in Table I, Table II and Table III, respectively. The tables show the symbol frequency spacings, normalized to Δf, for a modulation order up to $M = 8$.

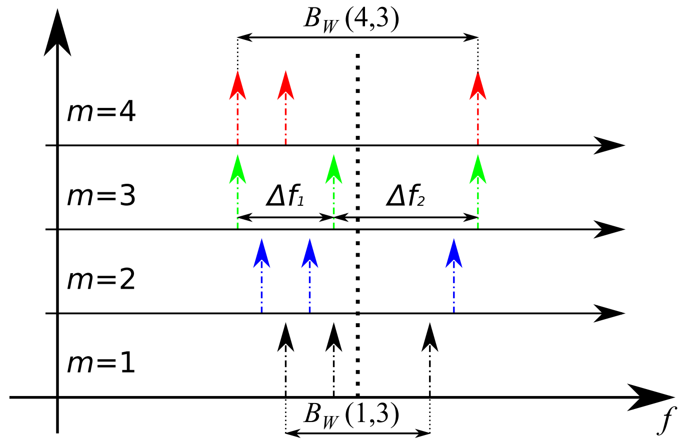

Fig. 5. Signal model illustrating the symbol construction for the proposed novel non-uniform FSK scheme, illustrated for 3 tones and M=4.

Moreover, higher orders are possible. These sets of frequency spacings result in sets of unique, non-coinciding IM_2 tones at the rectifier output, in baseband.

TABLE I
FREQUENCY SPACINGS ($\Delta f_i\,(m, N = 3)$) FOR NON-UNIFORM 3-TONE FSK, NORMALIZED TO Δf.

m	$\Delta f_1'\,(m, 3)$	$\Delta f_2'\,(m, 3)$
1	1	2
2	1	3
3	2	3
4	1	4
5	2	4
6	1	5
7	2	5
8	3	4
...		

TABLE II
FREQUENCY SPACINGS ($\Delta f_i\,(m, N = 4)$) FOR NON-UNIFORM 4-TONE FSK, NORMALIZED TO Δf.

m	$\Delta f_1'\,(m, 4)$	$\Delta f_2'\,(m, 4)$	$\Delta f_3'\,(m, 4)$
1	1	3	2
2	1	2	4
3	1	4	2
4	2	1	4
5	1	2	5
6	1	4	3
7	1	5	2
8	2	1	5
...			

The n-th tone's frequency for non-uniform multitone FSK can be found from these frequency spacings, using

$$f_{tone}\,(n, m, N, M, \Delta f) = f_c - \frac{B_W\,(m, N)}{2} + \sum_{i=1}^{n-1} \Delta f_i\,(m, N)\,, \quad \text{for } 1 \leq n \leq N, \quad \text{for } 1 \leq m \leq M. \tag{3}$$

with tone index n starting on the left of the spectrum, N the total number of tones, $\Delta f_i\,(m, N)$ the non-normalized

TABLE III
FREQUENCY SPACINGS ($\Delta f_i\,(m, N = 5)$) FOR NON-UNIFORM 5-TONE FSK, NORMALIZED TO Δf.

m	$\Delta f_1'\,(m,5)$	$\Delta f_2'\,(m,5)$	$\Delta f_3'\,(m,5)$	$\Delta f_4'\,(m,5)$
1	1	3	5	2
2	2	5	1	3
3	1	3	6	2
4	1	6	2	3
5	2	6	1	3
6	1	3	7	2
7	1	4	6	2
8	1	5	3	4
...				

frequency spacing, m the symbol index and $B_W\,(m, N)$ the symbol bandwidth, determined by

$$B_W\,(m, N) = \sum_{i=1}^{N-1} \Delta f_i\,(m, N)\,,\ \text{for } 1 \leq i \leq N-1. \quad (4)$$

These tone distributions result in $N\,(N-1)\,/2$ baseband tones. Of these, $N-1$ are located at frequencies equal to the frequency spacings $\Delta f_i\,(m, N)$. There are also additional intermodulating tones created at frequencies equal to

$$f_{IM}\,(m, N, i, k) = \sum_{j=i+1}^{k} \Delta f_j\,(m, N) + \Delta f_i\,(m, N)\,, \quad \text{for } 1 \leq i \leq N-1, \quad \text{for } i+1 \leq k \leq N-1. \quad (5)$$

The set of intermodulation products at frequencies, which equal the frequency spacings in Table I, Table II and Table III, and the additional intermodulation products determined by (5), are unique for each symbol. Hence, WIT is enabled by recognizing symbols at the rectifier output, by detecting the set of intermodulation products.

IV. ANALYSIS OF UNIFORM MULTITONE FSK

In this section a mathematical analysis of uniform multitone FSK is presented, leading to the expression for spectral efficiency. This is followed by experimental results, demonstrating the feasibility and limitations of the scheme.

A. Theoretical Analysis

After the signal is down-converted by the rectifier hardware, an FFT operation is performed. The only requirement to receive a signal modulated by the proposed uniform multitone FSK scheme, is to detect IM_2's frequency, which is known to be one of M values. Although baseband versions of the classical non-coherent FSK receivers can still be used, they are less flexible and the implementation in an integrated SWIPT receiver is less straightforward compared to an FFT-based approach. Hence, we elaborate on an FFT-based receiver. We investigate the inherent relation between throughput and bandwidth, extending this relation for two tones, which we already established in [18].

The bandwidth of the uniform FSK modulated multisine is defined as

$$B_W = (1 + (M-1)\,r)\,\Delta f\,(N-1) + 0.5r\Delta f. \quad (6)$$

The center frequencies of the FFT-bins f_{bin} are expressed by

$$f_{\text{bin}}(i) = \frac{F_{\text{s}}}{L} \times i, \quad \text{with } -\frac{L}{2} \leq i \leq \frac{L}{2}, \quad (7)$$

with i an integer bin-index, L is the number of samples in one symbol over which the FFT is taken and F_{s} is the sampling rate. The symbol size L is expressed in terms of the amount of baseband periods n in one symbol:

$$L = \frac{F_{\text{s}}}{\Delta f} \times n. \quad (8)$$

The width of one FFT frequency bin (B_w) can be found by combining (7) and (8) and equals

$$B_w = \frac{F_{\text{s}}}{\frac{nF_{\text{s}}}{\Delta f}} = \frac{\Delta f}{n}. \quad (9)$$

The FFT's frequency resolution has to be precise enough to ensure that the frequency shifts at the rectifier's output as described in section III, can be detected. This results in a constraint on the maximal bin-width, expressed by

$$B_w = \gcd(1, r)\Delta f, \quad (10)$$

with $\gcd(,)$ the greatest common divider operation. Since r can be rational, we assume that $\gcd(1, r = 0.5) = 0.5$ and $\gcd(1, r = 1.5) = 0.5$. The FFT bin-width determined by the amount of baseband periods per symbol is expressed by (9). The maximal bin-width determined by the detectable frequency shifts is expressed by (10). The relation between necessary amount of baseband periods per symbol and detectable frequency shift is found by combining these two equations, resulting in:

$$n_{min} = \frac{1}{\gcd(1, r)}. \quad (11)$$

This illustrates that the FFT has to be taken over more symbols to be precise enough to detect small frequency shifts, which will result in a relation between bandwidth and throughput as is shown next. The relation in (11) also ensures that the different locations for IM_2 align with the bins' center frequencies in (7).
The ideal information throughput is expressed by

$$T_P = \frac{\log_2(M)}{T_{\text{symb}}}, \quad (12)$$

where symbol duration $T_{\text{symb}} = n/\Delta f$. The throughput for the uniform scheme, in (12), is independent of N since the frequency spacings are uniform and dependent and do not encode individual information. This is different for the non-uniform scheme as will be shown in (15). The maximal throughput $T_{\text{P,max}}$ is found by substituting (11) in (12), resulting in

$$T_{\text{P,max}} = \gcd(1, r)\log_2(M)\,\Delta f. \quad (13)$$

The relation between maximal throughput and bandwidth is found by combining (6) and (13)

$$S_{\text{E,max}} = \frac{\gcd(1, r)\log_2(M)}{(1 + (M-1)\,r)\,(N-1) + 0.5r}. \quad (14)$$

By controlling the range of frequency spacings with M and r, it is possible to control throughput and power conversion efficiency. Fig. 6 confirms that $r = 1$ results in the largest possible throughput, for each modulation order and used bandwidth. The figure shows the relation between maximal throughput and bandwidth, both normalized to ΔF. A low two-tone FSK modulation order is shown to be beneficial when high spectral efficiency and low bandwidth is desired. Taking r=1 gives maximum throughput and spectral efficiency, however, r can be lowered to further trade-off spectral efficiency and power. This is because for small r the frequency spacing does not change a lot for different symbols, which leads to symbols with more power efficiency. This is because it is known that PCE increases with F_{ratio}, as will also be shown shown in Fig. 7. This is because have more power efficient symbols with tones at the spectrum edges. Fig. 6 also shows that r should be a rational number between zero and one, which is the considered region in the rest of this work. The data clearly indicates that the throughput for some $r > 1$ can also be achieved with lower bandwidth if $r < 1$. r should be rational because of the lcm operator in α in (10). The results for some rational $r < 1$ are indicated by dots in the figure. For non-rational $r < 1$, α would be very high, resulting in very low throughput, which is not shown in the figure because it is not an optimal parameter choice.

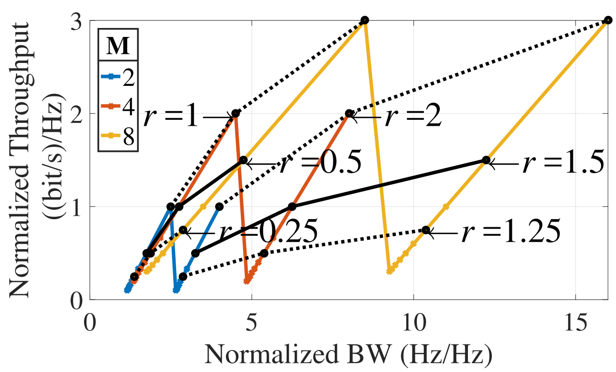

Fig. 6. Theoretical maximal throughput compared to bandwidth (B_W), normalized to ΔF for reasonable r, for an FFT-based detector and the proposed uniform FSK modulation scheme, with two tones.

B. Experimental Analysis

In this section we introduce the measurement results for uniform FSK modulation for SWIPT. Both power level and symbol rate are varied for different number of tones and modulation order.

Fig.7 shows the sub-optimal SER measurement results that we presented in [18], for varying modulation order M, frequency shift r and F_{ratio}. Measurements are taken over 10^3 and 10^4 symbols, depending on the required bandwidth and available hardware sample rate. The results show that SER increases for increasing F_{ratio}. SER increases more with higher F_{ratio}, for higher M and higher r. When using M=2 and r=0.25, SER is still below 10^{-4} for $F_{\text{ratio}} = 60$. Two-tone FSK with r=1 is shown here to enable a throughput of 5.79 Mbps, 6.95 Mbps and 5.78 Mbps at 10^{-3} SER for M=2, 4 and 8, respectively. This shows a five times increased throughput, compared to the biased ASK approach that was proposed in [14], [15], while enabling the use of two-tone signals for efficient WPT. Also, choosing r smaller than 1 lowers SER since it avoids that higher order intermodulation products fall inside the IM_2 bins which would confuse the detector in noisy situations and for high F_{ratio}. A low r also results in less difference in attenuation of the bins by the hardware since they are closer together. It is important to note that after improving

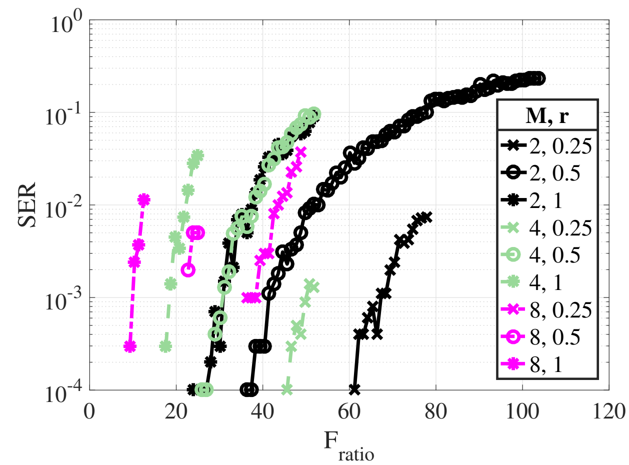

Fig. 7. Measurement results for SER, in terms of F_{ratio} with transmit power of 0 dBm, before improved signal processing , which theoretically removes the limiting F_{ratio}.

the signal processing in the receiver, the limits on F_{ratio} depicted in Fig.7 drastically increase. Because of transmitter hardware limitations, measurements were only possible up to $\Delta f = 60 * 10^5$ Hz. This corresponds to throughputs of 6 Mbps, 12 Mbps and 18 Mbps at 0 SER for M=2, 4 and 8, respectively. Hence, in a noiseless case, WIT bitrate using uniform frequency spacings is not limited by F_{ratio}. In addition, measurements for varying number of tones from 2 to 32 also show that this unlimited character is not affected by the number of tones, which is very beneficial for WPT. This conclusion is also confirmed by simulations using the model in Fig. 2. Both observations are major benefits over the biased ASK modulation scheme that we proposed earlier in [14], [15] where F_{ratio} was roughly limited to 20. However, increasing F_{ratio} leads to small output intermodulation products, which are very sensitive to noise. For a uniform two tone FSK signal with certain fixed noise level and $F_{\text{ratio}} = 3$, measurements show that $SER = 0.049, 0.135$ and 0.400 for $M = 2, 4$ and 8, respectively. This demonstrates that also these schemes are more sensitive to noise, with increasing modulation order.

The measured rectifier output voltage behavior for a two-tone in-phase input signal with varying frequency spacing is shown in Fig.8. These measurement results show that WPT benefits from a high symbol rate and frequency spacing, increasing mean output voltage. Hence, taking M=2 is optimal for WPT, considering a certain available bandwidth. WPT can further be optimized by lowering the frequency shift with r

to maximize Δf. However, this means that throughput will drop, as shown in Fig. 6. Increasing F_{ratio} also decreases the voltage swing, which is beneficial for a reliable output power, as shown in Fig. 8.

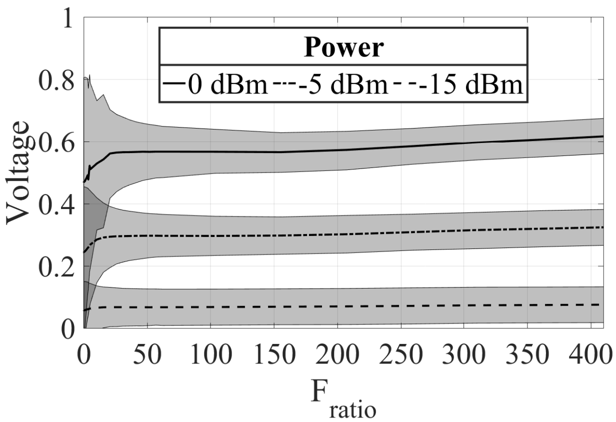

Fig. 8. Measurement results for rectifier output voltage swing (grey areas) and mean output voltage (solid lines) for a two-tone in-phase input signal, in terms of F_{ratio}, with transmit power of 0 dBm, -5 dBm and -15 dBm.

V. Analysis of Non-Uniform Multitone FSK

In this section a mathematical analysis of non-uniform multitone FSK is presented, leading to the expression for spectral efficiency. This is followed by experimental results, demonstrating the feasibility and limitations of the scheme.

A. Theoretical Analysis

The information throughput for non-uniform multitone FSK is expressed by

$$T_P = \frac{\log_2(M)(N-1)}{T_{\text{symb}}}, \tag{15}$$

where symbol duration $T_{\text{symb}} = n/\Delta f$, M the amount of symbols and N the number of tones per symbol. Here, the throughout does depend on the number of tones since N relates to the number of independent frequency spacings, which encode independent information. The maximal throughput $T_{\text{P,max}}$ is found by substituting (11) (with $r = 1$) in (15), resulting in

$$T_{\text{P,max}} = \log_2(M)(N-1)\Delta f. \tag{16}$$

The relation between maximal throughput and bandwidth depends on the number of symbols M and the number of tones per symbol N. Fig. 9 shows the bandwidth-throughput relation, normalized to Δf, depending on number of tones and modulation order, for non-uniform multitone FSK. WIT is optimized when choosing the parameter values that maximize spectral efficiency, the ratio of throughput over bandwidth. These points are indicated with black circles in the figure. The optimal modulation orders M are $[2]$, $[2, 4]$, $[4, 8, 16, 32, 64]$, $[32, 64, 128, 256, 512, 1024]$ and $[512, 1024, 2048]$ for 2, 3, 4, 5 and 6 tones respectively. Numerically optimizing spectral efficiency (S_E) results in the following relation between bandwidth from (4), frequency spacing and modulation order for the non-uniform scheme:

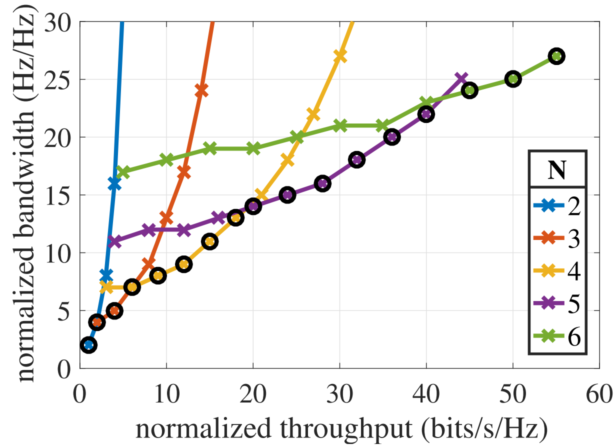

Fig. 9. Bandwidth-throughput relation for non uniform multitone FSK, both normalized to Δf, depending on number of tones and modulation order (from 2 to 2048, from left to right), with circles indicating the optimal selection of N and M concerning spectral efficiency.

$$\begin{aligned} S_{\text{E,max}} &= \frac{\log_2(M)(N-1)\Delta f}{B_W} \\ &= -0.137 * \ln(\log_2(M)) + 0.5135. \end{aligned} \tag{17}$$

This theoretical analysis of non-uniform multitone FSK and the analysis for uniform multitone FSK in section IV are summarized Fig. 10. The figure shows the trade-off between spectral efficiency, desired for WIT, and PAPR, desired for WPT. It is clear that non-uniform FSK results in better spectral efficiency whereas the uniform scheme fills the bandwidth more, resulting in higher PAPR. However, as will be shown in the next section, WIT using the uniform scheme can use a much higher F_{ratio}, improving WPT as well as throughput, which is certainly useful in some applications.

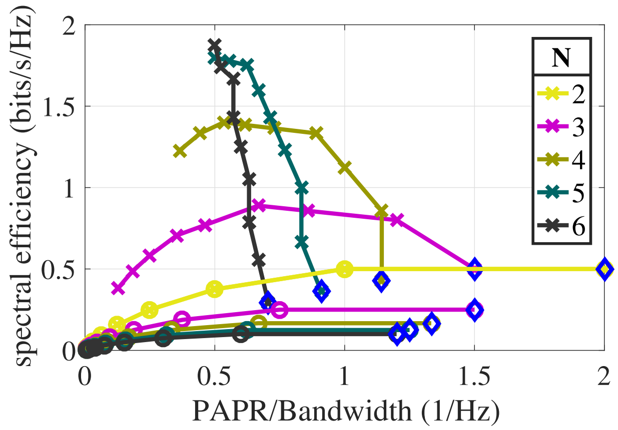

Fig. 10. Relation between spectral efficiency and PAPR for uniform (circle) and non-uniform (crossed) multitone FSK, for various number of tones and modulation orders with the $M = 2$ case indicated by diamonds.

B. Experimental Analysis

Fig. 11 shows the measured F_{ratio} upper-bound results for non-uniform FSK modulation, for our specific rectifier, while varying number of tones and modulation order. The results show that F_{ratio} needs to drop for increasing number of tones and increasing modulation order in order to still decode information. The main limiting factor is the non-uniform attenuation of the rectifier's low pass filter, leading to erroneous detection of the output set of intermodulation products. Hence we can choose between low symbol rate but high modulation order, or high symbol rate with low modulation order. Both impact throughput. This limit for the non-uniform scheme is similar to the limit found for biased ASK in [14], [15]. Contrary to the uniform scheme, F_{ratio} is limited for the non-uniform scheme and for (biased) ASK ([14], [15]), limiting throughput for a certain integrated rectifier receiver. However, (biased) ASK can be received more easily, without FFT, and the non-uniform scheme increases spectral efficiency, providing different trade-offs depending on what the application needs. The realizable throughput, calculated from Fig. 11 using (16), is shown in Fig. 12, showing that for our rectifier, the highest throughput can be reached by choosing a low modulation order and less than six tones.

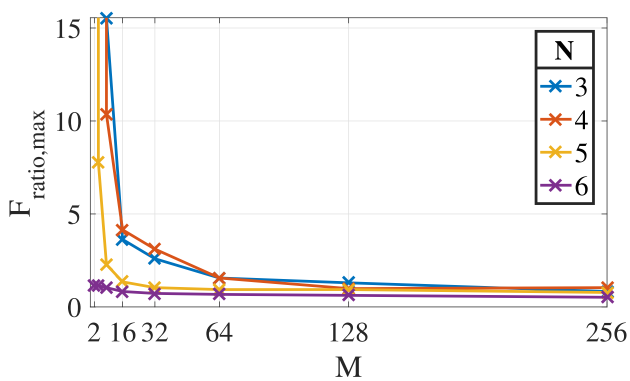

Fig. 11. Measurement results for non-uniform multitone FSK, demonstrating the upperbound on F_{ratio} for different number of tones N and modulation order M.

Fig. 12 shows the measured throughput upper-bound results for non-uniform FSK modulation, while varying number of tones and modulation order. It is well-known that the diode's

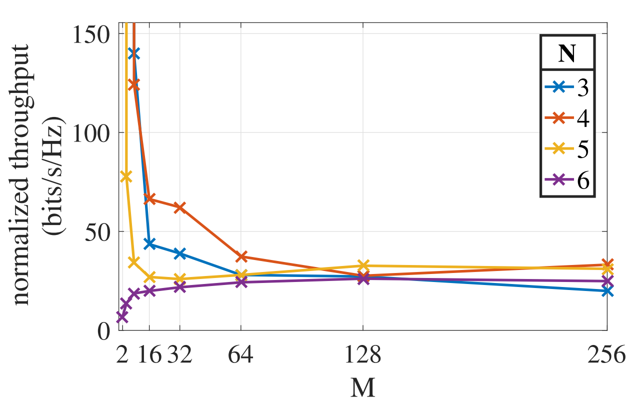

Fig. 12. Measurement results for non-uniform multitone FSK, demonstrating the upperbound on throughput for different number of tones N and modulation order M.

non-linear behavior is power dependent. Since our modulation schemes are based on that non-linearity, it is worth investigating whether their performance is power dependent as well. The measurement results in Fig. 13 demonstrate the impact of power on the non-uniform scheme's F_{ratio} upper-bound. The upper-bound decreases for lower power level. This means that the proposed modulation technique is more useful at high power levels. A limiting factor (on F_{ratio}) is the

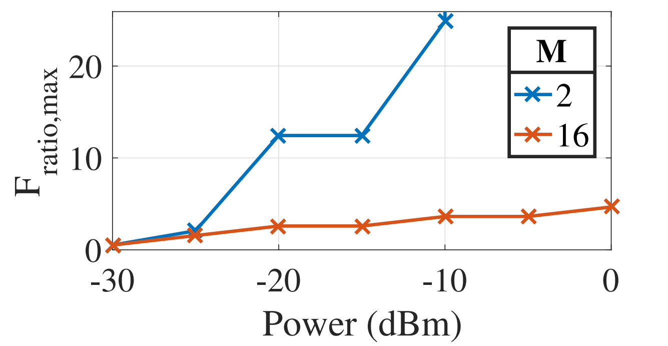

Fig. 13. Measurement results for impact of input power on F_{ratio} upper-bound, for a four tone non-uniform FSK modulated signal with varying modulation orders.

unfair attenuation of the rectifier 's low pass filter, causing errors in the bin-detection phase. As a solution, we propose an optimization in the next section.

VI. Optimization

The low pass filter at the diode's output attenuates the baseband frequency tones in a frequency dependent way. This results in a different attenuation for different symbols since the IM_2's are located elsewhere. This results in more errors for the symbols with high IM_2 frequency since they are attenuated more compared to the noise of intermodulation products in lower frequency bins. This is represented in Fig. 14 where each color represents a different symbol.

Measurement results presented earlier already showed that the uniform scheme is robust to this effect since only one baseband tone is detected, and no other strong baseband tones are present on the left of that tone. For the non-uniform scheme however, a lot of different baseband tones need to be detected, causing its sensitivity to this effect as presented earlier. Hence,

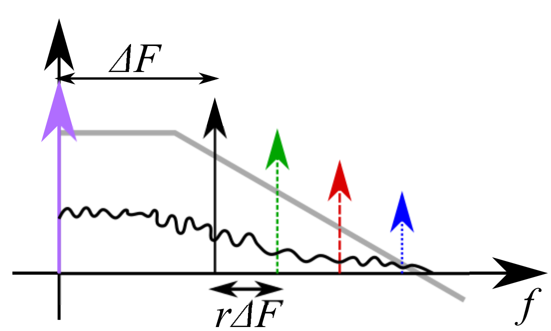

Fig. 14. Received signals in baseband, after passing the rectifier, with different symbol outputs represented by different styles.

a mathematical compensation is introduced to counter this effect. The ideal rectifier's low pass filter transfer function $H(i)$ is

$$H(i) = \frac{1}{1 + jF_{\text{ratio}}\left(1 + ri\right)}, \quad \text{for } 0 \leq i \leq M-1, \tag{18}$$

with i the symbol-index. A scaling factor for each bin is introduced in order to remove the unfair symbol attenuations before bin-detection is performed. This individual scaling is determined by the bin's attenuation relative to the first bin. The bin-scaling $K(i)$ is determined by

$$\frac{1}{K(i)} = \frac{|H(i)|}{|H(0)|} = \frac{\sqrt{1 + \left(F_{\text{ratio}}\right)^2}}{\sqrt{1 + \left(F_{\text{ratio}}\right)^2 \left(1 + ri\right)^2}}, \quad \text{for } 0 \leq i \leq M-1, \tag{19}$$

This mathematical compensation can be performed at the transmitter as precompensation of the signal, or at the receiver as postcompensation. Both are discussed next. Our experiments in the next sections show that the compensation factors of (19) are not optimal in practice since they are based on an independent and ideal low pass filter. This means that one should look for the optimal bin weights, starting from the theoretical factors in (19) for a certain operating point to fully counter the low pass filter effect.

Applying these bin weights can be done both at the transmitter or receiver side, which is called pre- and postcompensation respectively. Both are discussed next.

A. Postcompensation

When considering postcompensation, the transmitted tones of all symbols have equal amplitude and phase. At the rectifier output however, the low-pass filter attenuates intermodulation products with higher frequency spacing more. This leads to detection errors. This unfair attenuation can be compensated for by re-scaling the detected bins using (19). The measurement results in Fig. 15 demonstrate the huge benefit of applying post compensation, in case of a three tone non-uniform FSK modulated signal.

B. Precompensation

The low pass filter can also be countered at the transmitter, using precompensation. In this case, the weights of (19) are applied when the symbols are generated. This scaling of each symbols results in a varying amplitude of the RF signal. This causes an undesired voltage ripple at the diode's output. However, it allows the receiver to be less complex. The RF-signal's PAPR, in case of a uniform FSK modulation, is expressed by

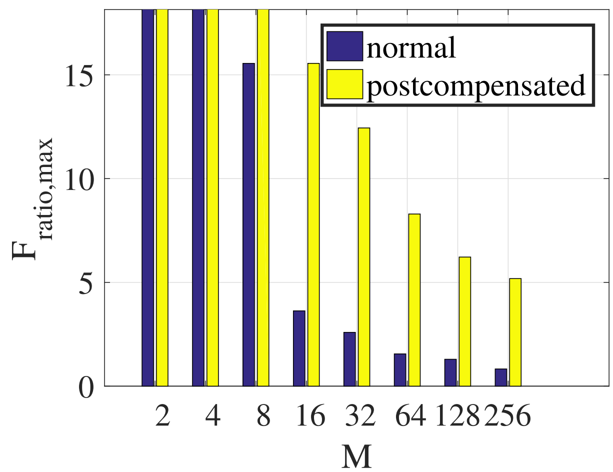

Fig. 15. Measurement results for F_{ratio} upper-bound, with and without post-compensation, for a three tone non-uniform FSK modulated signal with varying modulation orders.

$$\begin{aligned} &PAPR\left(N, M, F_{\text{ratio}}, r\right) \\ &= \frac{2N \times \dfrac{1 + F_{\text{ratio}}^2 \left(1 + r\left(M-1\right)\right)^2}{1 + F_{\text{ratio}}^2}}{\dfrac{M\left(F_{\text{ratio}}^2\left(\left(2M^2 - 3M + 1\right)r^2 + 6\left(M-1\right)r + 6\right) + 6\right)}{6M\left(F_{\text{ratio}}^2 + 1\right)}} \\ &= \frac{12N\left(1 + F_{\text{ratio}}^2\left(1 + r\left(M-1\right)\right)^2\right)}{\left(F_{\text{ratio}}^2\left(\left(2M^2 - 3M + 1\right)r^2 + 6\left(M-1\right)r + 6\right) + 6\right)}, \end{aligned} \tag{20}$$

where N is the number of tones used and M the modulation order. Since a standard multisine's PAPR is $2N$, (20) shows that PAPR increases.

VII. Conclusion

We propose two versions of a novel multitone FSK modulation technique for SWIPT, extending our proposed two-tone FSK modulation technique. This modulation scheme allows the information to be received by the rectifier hardware, without a power consuming local oscillator. Information is encoded in the frequency spacings of multitone signals allowing the usage of high PAPR signals for WPT, while transferring information as well. The proposed uniform multitone FSK technique fills the bandwidth with tones, optimizing PAPR and hence WPT. The proposed non-uniform multitone FSK modulation technique still uses multitone signals but with varying frequency spacing between tones to increase the amount of information per symbol, increasing spectral efficiency and WIT performance. We show that these kind of modulated signals allow to be downconverted using rectifier hardware, avoiding the need for a power consuming local oscillator. An optimization to counter the non-uniform low pass filter attenuation is proposed, increasing WIT even more in the non-uniform case. Our measurements show that both techniques perform well using a real power optimized rectifier. Both techniques allow to use high PAPR signals where each symbol carries both power and information, contrary to classical FSK and ASK. Our results show that the throughput of the uniform scheme is not limited by the rectifier hardware, contrary to (biased) ASK and the non-uniform scheme. However, (biased) ASK can be received more easily, without FFT and the non-uniform scheme increases spectral efficiency.

In the future, it might be interesting to analyze and compare the impact of frequency selective channels and corresponding precoders, as well as a typical mathematical rate-energy trade-off comparison between these unconventionel SWIPT modulation techniques (multitone FSK, biased ASK, two-tone ratio modulation and the PAPR modulation) and classical ASK and FSK.

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