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Original citation & hyperlink:
Haigh, P. A. , Le, S. T. , Zvanovec, S. , Ghassemlooy, Z. , Luo, P. , Xu, T. , Chvojka, P. , Kanesan, T. , Giacoumidis, E. , Canyelles-Pericas, P. , Le Minh, H. , Popoola, W. , Rajbhandari, S. , Papakonstantinou, I. and Darwazeh, I. (2015) Multi-band carrier-less amplitude and phase modulation for bandlimited visible light communications systems. IEEE Wireless Communications, volume 22 (2): 46 – 53
http://dx.doi.org/10.1109/MWC.2015.7096284

ISSN 1536-1284
DOI 10.1109/MWC.2015.7096284

Multi-band Carrier-less Amplitude and Phase Modulation for Bandlimited Visible Light Communications Systems

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***Abstract*—Visible light communications (VLC) is a technology with enormous potential for a wide range of applications within next generation transmission and broadcasting technologies. VLC offers simultaneous illumination and data communications by intensity modulating the optical power emitted by light-emitting diodes (LEDs) operating in the visible range of the electromagnetic spectrum (~370 – 780 nm). The major challenge in VLC systems to date has been in improving transmission speeds, considering the low bandwidths available with commercial LED devices. Thus, to improve the spectral usage, the research community has increasingly turned to advanced modulation formats such as orthogonal frequency division multiplexing (OFDM). In this article we introduce a new modulation scheme into the VLC domain; multi-band carrier-less amplitude and phase modulation (*m*-CAP) and describe in detail its performance in terms of bandlimited systems.**

I. Introduction

ADVANCED modulation formats are becoming increasingly important in visible light communications (VLC) systems [1-3] using light-emitting diode (LED) infrastructure. One of the primary aims of VLC is to provide high capacity broadcasting networks to next generation smart homes and offices, whilst simultaneously providing full room illumination. Therefore, one of the key challenges associated with VLC concerns the bandwidth limitation associated with commercial white LEDs, which behave as first order low-pass filters [4]. White light is produced using blue LEDs coated in yellow colour converting phosphor. Even though the blue LEDs offer bandwidths in the order of tens MHz, the colour converting phosphor substantially reduces this to just a few MHz, imposing a large impediment to achieving high speed networks. A straightforward method to combat the bandwidth limitation of LEDs is to employ an appropriate power pre-emphasis technique [1]. However, using this method will introduce additional signal distortion due to the nonlinear effects of the LEDs [5], limiting the achievable performance gain. As a result, achieving high capacities has been a major challenge for the VLC research community.

In recent years, orthogonal frequency division multiplexing (OFDM) has been the focus of enormous attention due to its ability to support spectrally efficient and high order modulation formats such as quadrature amplitude modulation (QAM) and efficiently overcome the distortion imposed by bandwidth limitation or non-flat fading. This enables an improvement in power and spectral usages over more traditional formats such as on-off keying (OOK), resulting in increased transmission speeds. For instance, transmission speeds in the region of 1 Gb/s/wavelength have been achieved in [2] by optimizing the OFDM modulation format using adaptive bit- and power-loading, while speeds in the order of hundreds of Mb/s have been achieved using OOK [4]. Adaptive loading techniques operate by adjusting the number of bits-per-symbol and power-per-subcarrier according to the measured error vector magnitude (EVM) at the receiver.

Regardless of the increased popularity of OFDM, the VLC research community is searching for alternative modulation formats. New multi-carrier techniques have been considered recently to save bandwidth in optical systems by compressing the sub-carrier spacing. Fast OFDM [6], spectrally efficient frequency division multiplexing [7] and faster than Nyquist [8] have all been studied and demonstrated for use in optical transmission systems, resulting in up to 50% bandwidth savings. However, complexity remains an issue that has to be resolved before these systems can be adapted to VLC.

Currently, one of the most novel and popular modulation formats is carrier-less amplitude and phase modulation (CAP), which is simple to implement in available DSP hardware and has been experimentally shown to offer improved transmission speeds in comparison to OFDM when utilizing the same physical link [2]. This is significant, as CAP modulation is straightforward to implement in real time, with the most

Manuscript received 1 September, 2014.

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computationally complex components being the single digital-to-analogue converter required, while the carrier frequencies are generated using finite impulse response (FIR) filters. On the other hand, OFDM requires the use of inverse and forward fast Fourier transforms which can be computationally expensive, depending on the required number of subcarriers.

Unfortunately, to the authors' knowledge, CAP has only been tested in the case of flat-band magnitude responses, which is a phenomenon that is seldom available in VLC, due to the low modulation bandwidths available from the LEDs; refer to Fig. 4(b) of [9]. Despite CAPs advantages, its candidature as a promising candidate for generic VLC systems is questionable. OFDM can be tailored to any frequency response using bit-loading algorithms, unlike CAP which must have a fixed modulation order cardinality. In [10] a solution to this problem was proposed for optical fibre links. The gross transmission bandwidth was divided into six sub-bands (or subcarriers, for consistency with OFDM nomenclature), relaxing the condition for a flat-band response and allowing the number of bits/symbol to be adjusted for each subcarrier.

Based on this, it follows that additional performance can be gained by increasing the number of subcarriers towards *m* (resulting in *m*-CAP), as the decreasing subcarrier bandwidths approximate towards flat-bands. On the other hand, by increasing to *m* subcarriers, the FIR filter requirement scales to 2*m* and consequently the relative computational complexity is increased accordingly. Simultaneously, [10] shows that with increasing *m*; the required sampling frequency approaches the Nyquist rate, which is a significant and unexpected advantage. In Fig. 1 the concept of *m*-CAP is illustrated with $m = 1, 2, 10$. By increasing *m*, the subcarrier bandwidth is reduced, offering additional protection to attenuation caused by the low-pass LEDs in comparison to the high bandwidth subcarriers (i.e. low *m*). Therefore a trade-off between complexity and performance will arise just as with OFDM systems. The objective of this article is not to analyse the relative computational complexities between OFDM and *m*-CAP and hence this discussion is left open for research. In this article, we propose *m*-CAP as a standalone modulation format for VLC and consider its performance in LED bandlimited systems. The proposed system is studied using realistic system and device models and verified through numerical simulations.

II. *m*-CAP MODULATION FOR VLC

A block diagram for an *m*-CAP VLC system is illustrated in Fig. 2.

A. *m-CAP Generation*

In this section a description on *m*-CAP generation is provided. First, *m* independent data streams D_m are generated and mapped into the QAM constellation (Fig. 2(a)). Here 16-QAM will be considered to illustrate the concept. The choice of constellation cardinality is insignificant as without any loss of generality, the ensuing discussion is valid across every order of QAM. The data is upsampled by means of zero padding by a number of samples per symbol N_{ss} in order to match the system sampling frequency before the real $\Re$ and imaginary $\Im$ components of the modulated data are isolated. The components are passed through real and imaginary transmit filters whose impulse responses form a Hilbert pair [10]. The impulses are found as the product of a cosine or sine (real or imaginary, respectively) with root-raised cosine filter (RRCF) as shown in Fig. 2(b), while the mathematics can be found in [11]). The minimum carrier frequency must be at least twice the RRCF pulse width. Clearly the carrier frequency of the subcarrier is controlled by the frequency of the (co)sine wave and hence is introduced by the FIR filters, unlike QAM which requires a local oscillator. This is the main difference between the two formats and is a significant advantage of CAP, since incoherent time-reversed matched filters can be deployed at the receiver, eliminating the use of a complex phase locking circuit. The real and imaginary components are added before transmission via intensity modulation of the LED.

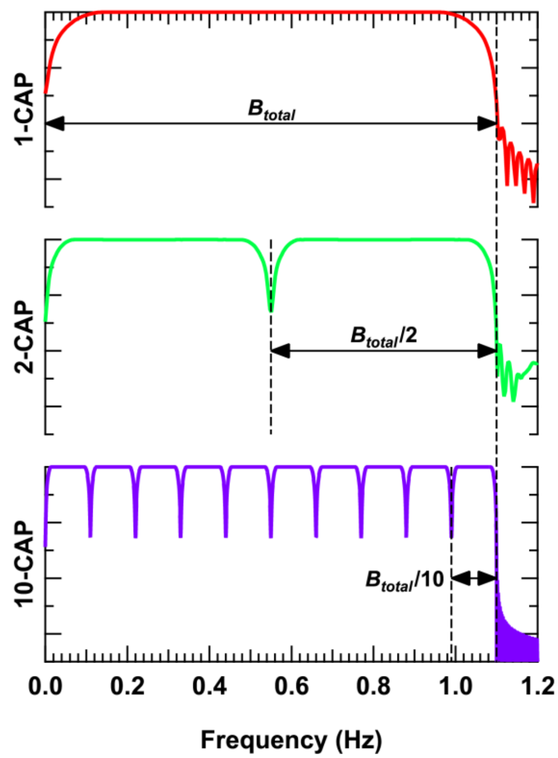

Fig. 1 The concept of *m*-CAP illustrated in terms of the frequency response for several values of *m*. It should be noted that B_{total} indicates the system bandwidth as will be discussed

The required sampling frequency and number of samples-per-symbol can be referred to mathematically in the literature and hence are not shown here [10]. Considering the use of a RRCF, the minimum subcarrier bandwidth is controlled by the roll-off factor, β, as expected, which varies between $0 \leq \beta \leq 1$. A larger β means a $1 + \beta$ greater bandwidth requirement as is illustrated in Fig. 3 for a normalised signal bandwidth. In this work we have set $\beta = 0.1$ due to (*i*) consistency with recent literature [10] and (*ii*) a lower β means a higher spectral efficiency can be achieved due to more effective spectral usage.

Further parameters adopted in this article as follows; the overall system baud rate is set to unity, meaning that each subcarrier has a baud rate of $1/m$. This condition ensures that the bandwidth remains constant regardless of *m* which is important when considering a bandlimited link as will be described in the following sentences. In order to examine the performance of *m*-CAP in terms of a bandlimited VLC system, two approaches can be adopted:

1. Maintaining a constant -3 dB modulation bandwidth and increase the overall baud rate.
2. Maintaining a constant overall unity baud rate and vary the LED low-pass cut-off frequency.

In this article we adopt the latter approach for simplicity, and hence the overall baud rate is selected as unity throughout. Therefore, recalling the selection of β = 0.1, the total bandwidth requirement is calculated as 1.1 Hz as is shown in **Fig. 4**.

In general, LEDs exhibit a non-linear electro-optic response (i.e. current to optical power) [12], however this varies from device to device and from one model to the next [12]. For instance, polymer LEDs can display excellent linearity [9], while conventional devices can heavily distort the transmit signal [12] and thus a generalized solution is impossible to consider at this stage. As a result we select a perfectly linear electro-optic response in this article as it is not our intention to discuss this type of signal distortion concurrently with bandwidth distortion at this stage. If the non-linear properties were taken into account, the *m*-CAP system would still work, but may require customisation to the specific LED chosen. The signal propagates over the line-of-sight indoor channel (reported mathematically in [9]), excluding multi-path components, and is detected by an ideal positive-intrinsic-negative (p-i-n) photodiode.

B. Modulation Bandwidths

As reported in [4], the major impediment to achieving high transmission speeds in VLC is the slow temporal response of the yellow phosphor, limiting bandwidths to a few MHz. This means that achieving high transmission speeds is a considerable challenge facing the VLC research community. It is well known that the frequency response of LEDs can be modelled, as a good first approximation, by a first order low-

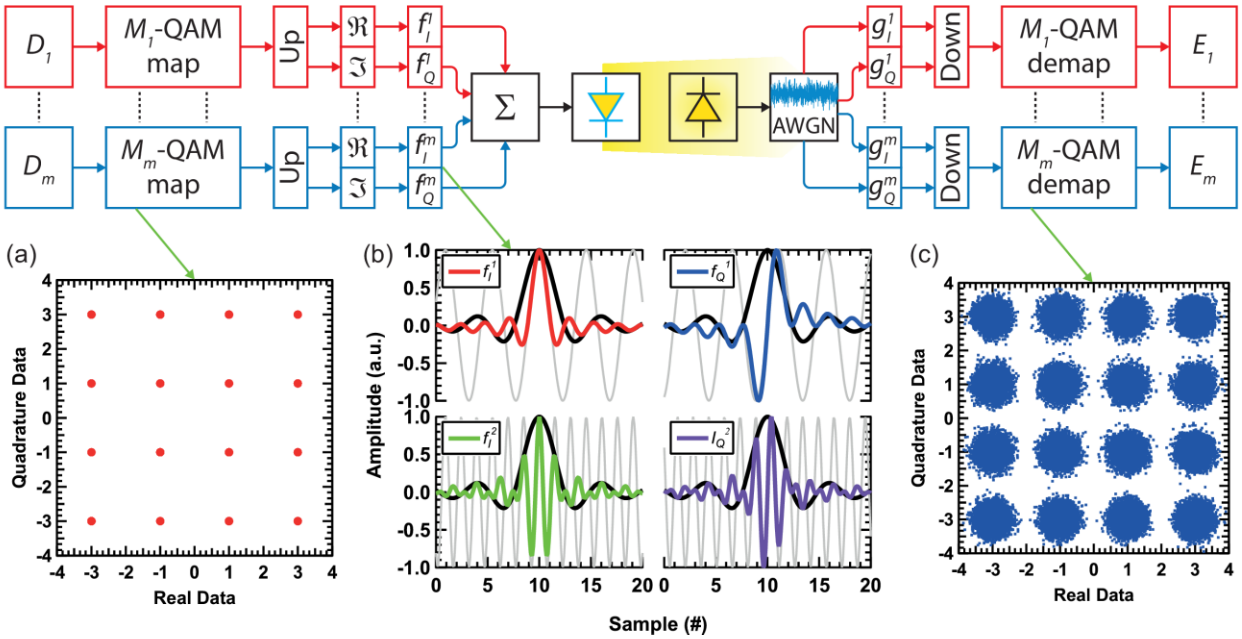

Fig. 2 The schematic block diagram of the proposed system. It should be noted that in (b) the impulse responses displayed are for m = 1 and 2 at the minimum carrier frequencies and variations on these pulse shapes are likely, depending on the desired carrier frequency <<change figure to remove f/g notation>>

pass filter [4], as the parasitic components will vary from device-to-device. For example polymer LEDs have substantially more parasitics (i.e. capacitance) than conventional inorganic devices. The LED bandwidths f_c under test are selected as a fraction of the overall 1.1 Hz signal bandwidth as follows: $f_c = \{0.1, 0.25, 0.5, 0.75, 0.9\}$. As an example, this means that when f_c = 0.5 the cut-off frequency of the LED is given by the ratio of the signal bandwidth to the LED bandwidth, i.e. 1.1/0.5 = 0.55 Hz, as illustrated in **Fig. 4** for 5-CAP, 10-CAP and 20-CAP.

It is clear from **Fig. 4** that as the order of *m* increases (i.e. a higher number of subcarriers); each subcarrier suffers reduced distortion due to the attenuation caused by the low pass filter in spite of the fact that the envelopes of each signal possess the same shape. The reason for this is that each subcarrier has a lower bandwidth requirement for higher orders of *m*-CAP and hence is less prone to distortion caused by attenuation. It should be noted that the noise performance will not be improved as the total signal power for every value of *m* is equal. Thus, it would be expected that additional performance will be obtained when using higher values of *m* in bandlimited links, at the cost of additional hardware complexity via the increased requirement for FIR filters.

C. Receiver

At the receiver the sampling frequency should be equal to that at the transmitter, and hence needs to be set as described in [10], which shows that with an increasing value of *m*, the sampling frequency approaches the Nyquist limit. The implication of this is that when using *m*-CAP, transmission speeds at least equal to those available in conventional 1-CAP will be available whilst using a lower sampling frequency that approaches the Nyquist limit, thus relaxing the conditions for high speed DACs, which saves cost and complexity.

The dominant noise source present in the system is introduced by the receiver electronics and is modelled as additive white Gaussian noise (AWGN). Once more referring to Fig. 2, the received signal is split into 2*m* branches (real and imaginary for each *m*) and passed through reverse-time real and imaginary receiver filters matched to the transmit filters before down-sampling and constellation de-mapping (Fig. 2(c)) in order to provide an estimate of the symbols E_m for comparison with D_m in a symbol-by-symbol bit error rate (BER) measurement.

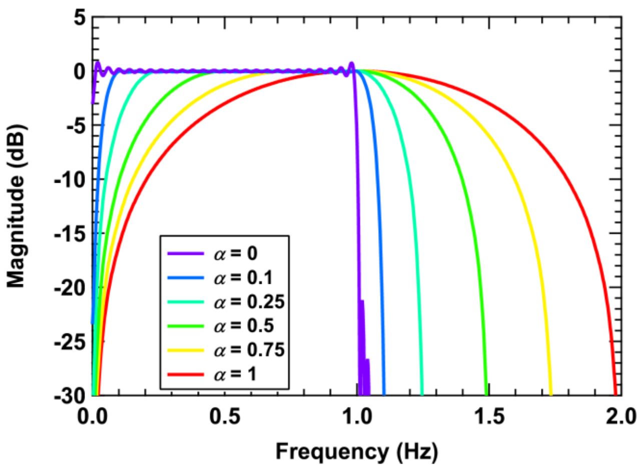

Fig. 3 Possible 1-CAP pulse shapes considering a varying β

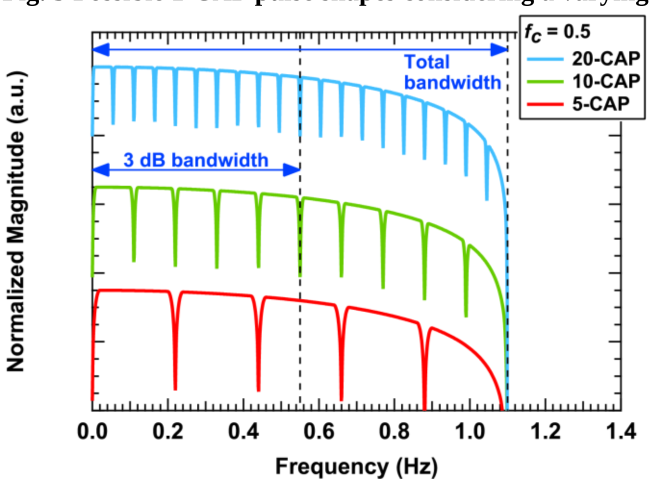

Fig. 4 Example frequency responses considering f_c = 0.5; note that the distortion on each subcarrier for high values of *m* is less than that for lower values of *m*

D. FIR Filters

Several advantages can be gained by dividing the desired transmission bandwidth into *m* subcarriers. The most important of which is that the attenuation observed over the signal frequencies due to the low pass filter is severely reduced due to the smaller subcarrier bandwidth requirements, as mentioned in the previous section. Therefore, here we examine the error performance across five different values of *m*, namely $m = \{1, 2, 5, 10, 20\}$, chosen to allow a large window of observation into any improvement. It should be noted that $m = 20$ corresponds to 40 FIR filters in the transmitter alone, and a further 40 in the receiver, amongst other resources. Therefore, the length of the filters is crucial; and hence studies have been performed that investigate the effect on varying lengths [13] and demonstrate improved BER performance with longer filter lengths. Depending on system requirements such as the target bit rate, BER or application, amongst others, forty FIR filters could be considered excessive, depending on the available resources. Therefore, it is necessary to keep in mind this trade-off between performance improvement and implementation complexity with *m*-CAP systems.

III. *m*-CAP Bit Error Rate Performance

A. Simple bandlimited m-CAP

The BER performance of the proposed bandlimited *m*-CAP systems are shown for f_c = 0.5 in Fig. 5(a)–(d), which show the performance of each subcarrier (denoted as *s* in Fig. 5) starting with 20-CAP in Fig. 5(a), 10-CAP in Fig. 5(b), 5-CAP in Fig. 5(c) and finally 2-CAP and 1-CAP, both in Fig. 5(d). The BER is presented as a function of the energy-per-bit to noise power spectral density ratio (E_b/N_o). The BER target is selected to be 10^{-3}, which is slightly lower than the International Telecommunication Union's (ITU's) recommended error floor (3.8×10^{-3}) when using a forward error correction (FEC) with overhead of 7% [14].

Fig. 5(a) illustrates that the BER performance of the vast majority of subcarriers approaches the theoretical BER limit for 16-QAM systems, with only a few subcarriers (s = {19, 20}) failing to converge to the theoretical limit. This is attributed to the following reasons:

1. The limited frequency response of the LED means the first half of the subcarriers s = 1:10 are not affected by the attenuation introduced by the LED low-pass filter; they are within the modulation bandwidth and hence have a maximum signal attenuation of 3 dB at s = 10.
2. The second half of the subcarriers, s = 11:20 incurs an additional E_b/N_o power penalty of several dB (approximately ~1 – 3 dB) in the best case, and failure in the worst cases, as can be seen for s = 19 and 20.
3. In the worst case this is because an ideal low-pass filter is used and hence a roll-off of 20 dB per decade is expected. So between s = 10 and s = 20, the attenuation should be roughly 20 dB and hence significantly degraded performance is shown at s = 20. This is also reflected in the constellation diagrams shown inset at E_b/N_o = 11 dB, for cases s = 1, 18, 19 and 20, showing a continual degradation.

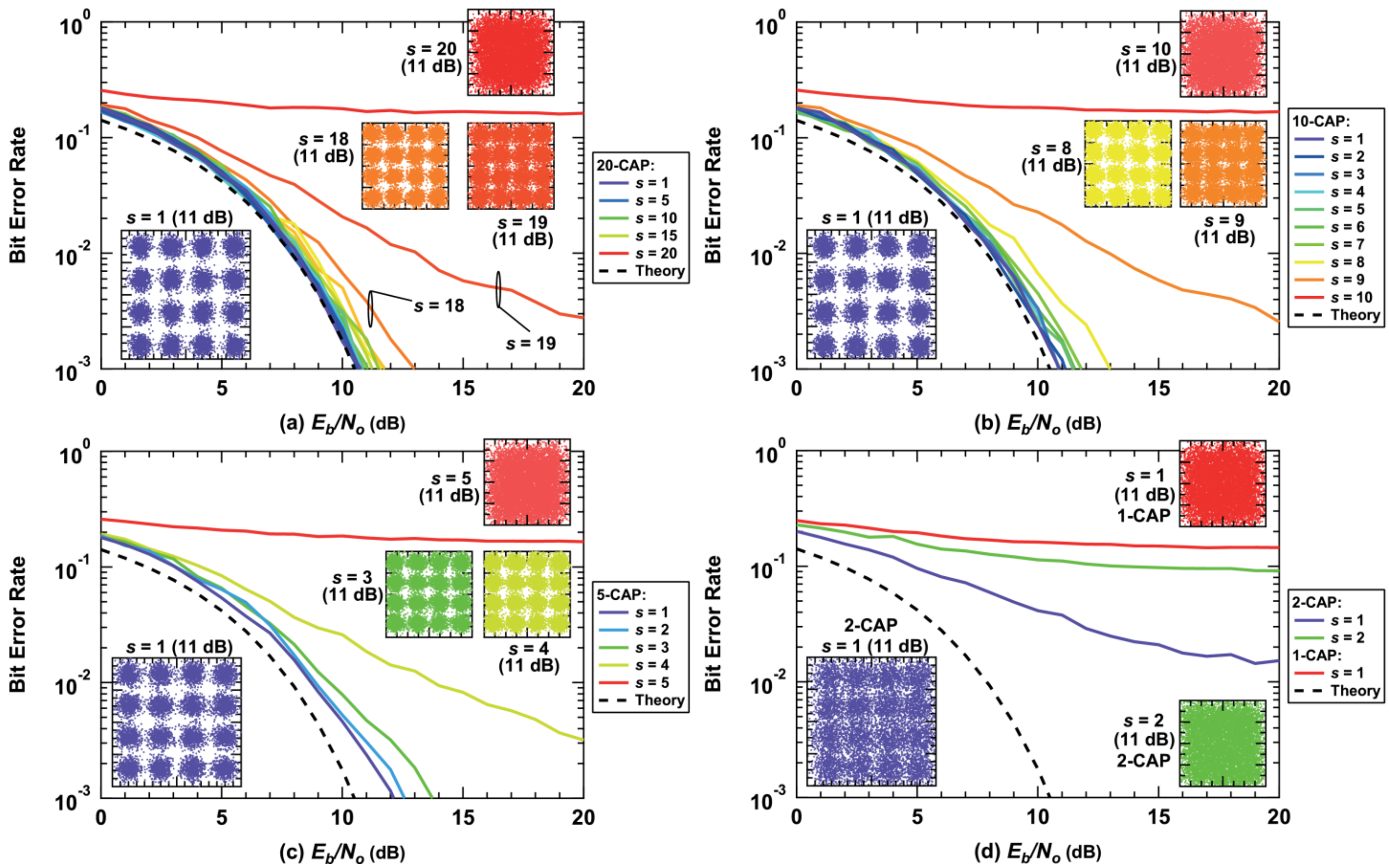

Fig. 5 BER performance as a function of SNR for *m* = (a) 20, (b) 10, (c) 5 and (d) {2, 1}. Clearly the BER performance improves with increasing *m* as is reflected in the constellation diagrams all of which consider a SNR of 11 dB

Fig. 5(b) illustrates the BER performance of a 10-CAP link under the same operating conditions of $f_c = 0.5$. A very similar profile to 20-CAP can be observed, however some very discrete differences between the two cases can be found; the most important is a slight power penalty for the best performing subcarriers $s = 1, 2, \ldots, 8$ of 0.5 to 2.5 dB in comparison to the theoretical performance of 16-QAM. Recalling that the 10-CAP subcarrier bandwidth requirement is twice that of the 20-CAP case, the attenuation of the low-pass filter can be seen as the cause for this slight power penalty, since a larger attenuation acts on the subcarriers due to the additional bandwidth requirements. This is reflected in the constellation diagrams shown inset for $s = 1, 8, 9$ and 10 for $E_b/N_o = 11$ dB.

Next, in Fig. 5(c), the same effect as the previous cases can be observed in 5-CAP with an exaggerated power penalty in comparison to the 16-QAM theoretical case which has increased to ~1.5 dB in the best case ($s = 1$) and 3.5 dB in the worst ($s = 3$). Larger subcarrier values fail due to the large attenuation observed as expected and once more this is reflected in the constellation diagrams inset. Finally, the results are essentially confirmed Fig. 5(d), where both 2-CAP and 1-CAP fail to meet the BER target in the E_b/N_o range considered. It should be noted that the remaining f_c values do follow the same trend but are not reported here due to space considerations.

B. Adaptive Bit-Loading within Bandlimited m-CAP

One advantage of *m*-CAP that has not been explored so far in this article is the possibility to change adaptively the order of the modulation format to suit a given system, based upon the measured signal to noise (SNR) values at the receiver. This technique is called adaptive bit-loading and is well known and widely reported in OFDM technologies [2]. To provide a brief overview, the same block diagram as proposed in Fig. 2 can be implemented but the process starts with the lowest modulation constellation/format of binary phase shift keying (BPSK). The principles of transmission are exactly the same as described previously, however at the receiver a BER measurement is not made in a symbol-by-symbol manner as has been done so far in this work. Instead, the root-mean square EVM of the received BPSK constellation is measured and the SNR is extracted based on a relationship that can be referred to in [15]. The measured SNR values are compared with the known SNR values, considering a target BER, for every order of *M*-QAM and the maximum possible number of bits/symbol are loaded into each subcarrier in order to improve the spectral efficiency of the system, and avoid errors in subcarriers where the SNR is insufficient. As always; this comes as a trade-off with hardware complexity as the adaptive algorithms will require resources within both the transmitter and receiver.

We implemented such an adaptive *m*-CAP system based upon the adaptive bit-loading technique for SNRs of 10, 20 and 30 dB for every value of f_c and the results are illustrated in

Fig. 6. For the system exhibiting the high SNR of 30 dB, we found that spectral efficiencies (b/s/Hz) can approach slightly short of 10 b/s/Hz in the case of f_c = 0.9, as the vast majority of subcarrier can be loaded with higher order values of *M*-QAM, thus allowing a high throughput link to be supported. As expected, the spectral efficiencies decrease for decreasing values of f_c. The decrease is approximately linear across the range of *m*. It is clearly indicated in the 30 dB case, that high spectral efficiencies between ~8 – 10 b/s/Hz can be achieved when using $m \rightarrow 20$. However, the improvement shown for lower values of *m* could be considered marginal when reflecting on the additional hardware complexity that arises from high values of *m*. For example *m* = 10 can achieve spectral efficiencies within the range of ~9 – 6 b/s/Hz for f_c = 0.9 – 0.1, respectively and it appears that an asymptotic level can be found in each case, where limited improvement is available, even as *m* tends to a very large number. This is supported by the 20 dB case, which shows a similar trend. The spectral efficiency is much more consistent with less variation over a wider range of *m*, implying that the asymptote is reached earlier. Finally for the 10 dB case due to noise restrictions no improvement in spectral efficiency can be found for any value of *m*.

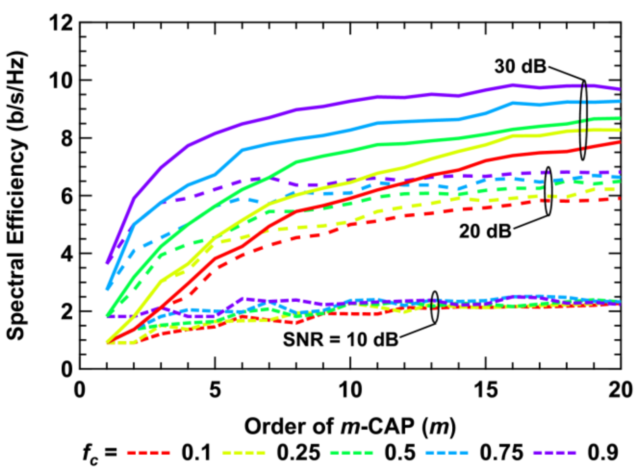

Fig. 6 Spectral efficiencies of the *m*-CAP systems considering five degrees of bandlimitation; spectral efficiencies up to ~ 10 b/s/Hz can be achieved with sufficient SNR at a BER of 10^{-3} in this case

IV. Future Outlook and Conclusions

Here we have shown the considerable potential for *m*-CAP systems in a bandlimited VLC environment. We have shown high spectral efficiencies can be achieved even in highly bandwidth limited environments, depending on the received SNR. However, the only way to achieve this performance is to use the adaptive bit-loading techniques as outlined here.

If *m*-CAP were to be adopted in VLC systems, further investigation is required. Here we have provided the first deep insight to adopt this advanced modulation format and illustrated the potential improvement attainable using this scheme. The next stage in the theoretical work is to examine the impact on the link considering the LED non-linearity. In parallel to the enormous potential outlined here by simulations we are finalizing an experimental test-bed to support our simulations, since there is currently no evidence in the literature that reports on an experimental *m*-CAP VLC link. It would be of the upmost importance in the next step to examine the performance of 1-CAP and *m*-CAP in terms of BER performance for a series of experimental measurements.

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