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Experimental Demonstration of High-Speed 4×4 Imaging Multi-CAP MIMO Visible Light Communications

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***Abstract*—In general, visible light communication (VLC) systems, which utilise white light-emitting diodes (LEDs), only offer a bandwidth limited to the low MHz region. Therefore, providing VLC-based high data rate communications systems using VLC becomes a challenging task. To address this challenge, we propose a solution based on multiplexing in both the frequency and space domains. We experimentally demonstrate a 4×4 imaging multiple-input multiple-output (MIMO) VLC system (i.e., space multiplexing) utilising multi-band carrier-less amplitude and phase (m-CAP) modulation (i.e., frequency multiplexing). Independently, both MIMO and m-CAP have separately shown the remarkable ability to improve the transmission speeds in VLC systems, and hence, here we combine them to further improve the net data rate. We investigate the link performance by varying the number of subcarriers m, link distance L, and signal bandwidth B_{sig}. From all the values tested, we show a data rate of $\sim$249 Mb/s can be maximally achieved for m = 20, B_{sig} = 20 MHz, and L = 1 m, at a bit error rate of 3.2×10^{-3} using LEDs with $\sim$4 MHz bandwidth.**

***Index Terms*—Modulation bandwidth, multi-band carrier-less amplitude and phase modulation, multiple-input multiple-output, visible light communications.**

I. Introduction

WIRELESS access technologies have been continuously evolved in response to ever increasing capacity demands resulting from the wide spread use of smart and mobile devices. Current 4th generation wireless systems with a limited frequency spectrum will not be able to apportion sufficient bandwidth to cope with the exponentially growing traffic, and hence, highly spectrally efficient communication techniques are one of the key considerations in research towards the 5th generation networks [1]. Within this context, visible light communications (VLC) can be used as a complementary wireless technology [2] to radio frequency based schemes mostly in applications where high bandwidth B and low latencies are the main requirements. VLC has received growing research interest over the last decade; mostly due to the widespread use of light emitting diodes (LEDs) in solid-state lighting systems [3]. However, VLC systems have a number of challenges including (*i*) LEDs with limited bandwidths B_{LED} (typically $<$ 5 MHz, or higher for micro-LEDs); (*ii*) LED non-linearity; (*iii*) multipath induced inter-symbol interference (ISI); (*iv*) blocking and shadowing; and (*v*) limited mobility [3], [4], which limits the maximum achievable data rate R_b within a typical indoor environment.

Numerous of schemes have been proposed to overcome these limitations [5]–[10]. As for the B_{LED} and ISI limitations, the most widely adopted schemes include pre and post equalization, blue filtering, multiplexing and parallel transmission, and spectrally efficient modulation as well as multi-carrier transmission [11]–[13]. Of the latter, orthogonal frequency division multiplexing (OFDM) offers spectral efficiency only through compatibility with bit- and power-loading algorithms, thus enabling a higher number of bits/symbol/subcarrier [14], [15]. In [16]–[18] OFDM VLC links with R_b of several Gb/s were reported, and in [19] the aggregate R_b was increased to 3.4 Gb/s using a combination of OFDM and wavelength division multiplexing (WDM). Despite many advantages that OFDM offers, there are a number of drawbacks including (*i*) limited dynamic range, (*ii*) relatively high peak-to-average power ratio due to the LED nonlinearity; (*iii*) sensitivity to the carrier offset and drift, and phase noise; which result in lower overall spectral efficiency [20]–[22]

Alternatively, carrier-less amplitude and phase (CAP) modulation was shown to outperform OFDM in terms of R_b over the same transmission span L [23]. In [24], a CAP-VLC system employing a red-green-blue (RGB) LED and a hybrid post-equalizer was reported with R_b of 3.22 Gb/s. Overall, the CAP system based on intensity modulation and direct detection is less complex with improved performance compared to OFDM, which allows relatively higher R_b using optical and electrical components with limited B [25]. However, Despite superior R_b and bit error rate (BER) performance in comparison to OFDM, the major impediment in implementation of CAP VLC is that it is very sensitive to the high frequency attenuation [26], which is prevalent in VLC systems due to the LED frequency response (i.e., a low pass filter (LPF)). In [27] multi-band CAP (m-CAP) was first proposed for optical

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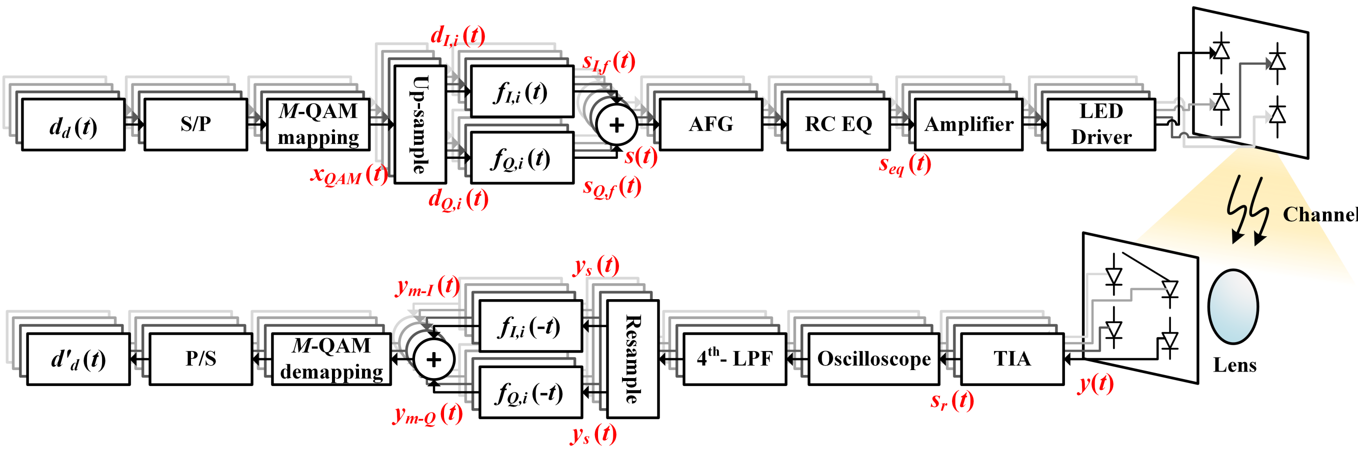

Fig. 1. Schematic block diagram of the 4 × 4 imaging MIMO m-CAP system

fibre communications before adoption by VLC community in [26], where the system B was evenly divided over m = 10 subcarriers, each employing CAP with successively higher carrier frequencies over a low modulation bandwidth. It was found that a higher signal-to-noise ratio (SNR) per subcarrier could be supported due to the alleviation of the band-limiting conditions. Furthermore, the implementation of a bit-loading algorithm, which is not possible in 1-CAP, enabled the demonstration of a record (at the time of writing) spectral efficiency η_s of 4.85 b/s/Hz.

Simultaneously, the multiple-input multiple-output (MIMO) scheme can be adopted to increase the R_b. In indoor environments multiple LED-based lighting fixtures are used to provide sufficient illumination levels (i.e., 200-1000 lux [28]), which facilitates the implementation of MIMO-VLC [29]. An imaging MIMO-VLC system using micro-LEDs ($B_{\text{LED}} > 20$ MHz) and R_b of 920 Mb/s over L of 1 m was experimentally demonstrated in [30]. In [31], an experimental 4 × 4 optical non-imaging MIMO-VLC link with an aggregate R_b of 50 Mb/s over L of 2 m was demonstrated. In this work, for the first time, we experimentally demonstrate a 4 × 4 imaging MIMO-VLC system based on m-CAP (i.e., a combination of frequency- and space-division multiplexing), and investigate the link's BER performance as a function of m (up to 20), a range of signal bandwidth B_{sig} , and L. We also examine the net data rate R_{net} and the net spectral efficiency η_{net}; which shows that for a given m and L, increasing B_{sig} leads to an increase in R_{net}, and reduced η_{net}. We demonstrate a maximum R_b of ∼249 Mb/s at a BER of 3.2×10^{-3} for m = 20, B_{sig} = 20 MHz and L = 1 m.

The rest of this paper is organised as follows: in Section II the system setup is described; results and discussions are shown in Section III, and finally, conclusions are drawn in Section IV.

II. System Setup

As a proof of concept, Fig. 1 shows the schematic block diagram of the proposed experimental system, where m independent pseudorandom binary sequences (PRBS) $d_d(t)$ in the non-return to zero (NRZ) format of length $2^{17} - 1$ are generated in the MATLAB domain. Following serial-to-parallel conversion (S/P), $d_d(t)$ is then mapped onto the M-ary quadrature amplitude modulation (M-QAM) constellation symbols $x_{QAM}(t)$, where the cardinality is given by $M = 2^b$ and b represents the number of bits/symbol. $x_{QAM}(t)$ is then up-sampled by the number of samples/symbol n_{samp} as given by [27]:

$$n_{\text{samp}} \;=\; > \; 2m(1+\beta), \tag{1}$$

where m represents the number of subcarriers, and β is the roll-off factor of the square root raised cosine (SRRC) filter, which is set to 0.15 for consistency with the literature [26], [27].

Both up-sampled signals $d_{I,i}(t)$ and $d_{Q,i}(t)$ of the i^{th} subcarrier are then passed through the real (I) and imaginary (Q) SRRC filters the outputs of which are given as [26]:

$$S_{I,f,i}(t) \;=\; d_{I,i}(t) \otimes f_{I,i}(t), \tag{2}$$

$$S_{Q,f,i}(t) \;=\; d_{Q,i}(t) \otimes f_{Q,i}(t), \tag{3}$$

where $\otimes$ denotes time domain convolution, $f_{I,i}(t)$ and $f_{Q,i}(t)$ denote the impulse response of the I and Q SRRC filters of the i^{th} subcarrier, respectively, which are given by [32]:

$$f_{I,i}(t) \;=\; \cos(2\pi f_{c,i} t). \left[\frac{\sin\left(\frac{\pi t}{T_s}(1-\beta)\right) + 4\beta\frac{t}{T_s}\cos\left(\frac{\pi t}{T_s}(1+\beta)\right)}{\frac{\pi t}{T_s}\left(1-\left(4\beta\frac{t}{T_s}\right)^2\right)}\right], \tag{4}$$

$$f_{Q,i}(t) \;=\; \sin(2\pi f_{c,i} t). \left[\frac{\sin\left(\frac{\pi t}{T_s}(1-\beta)\right) + 4\beta\frac{t}{T_s}\cos\left(\frac{\pi t}{T_s}(1+\beta)\right)}{\frac{\pi t}{T_s}\left(1-\left(4\beta\frac{t}{T_s}\right)^2\right)}\right], \tag{5}$$

where T_s is the symbol duration. The carrier frequency for the i^{th} subcarrier is given by [33]:

$$f_{c,i} \;=\; \frac{(2i-1)B_{sig}}{2m}, \tag{6}$$

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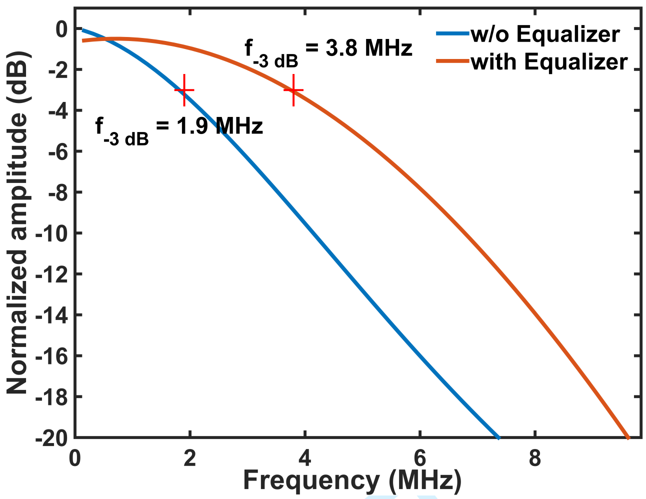


Fig. 2. LED frequency response with and without an equalizer

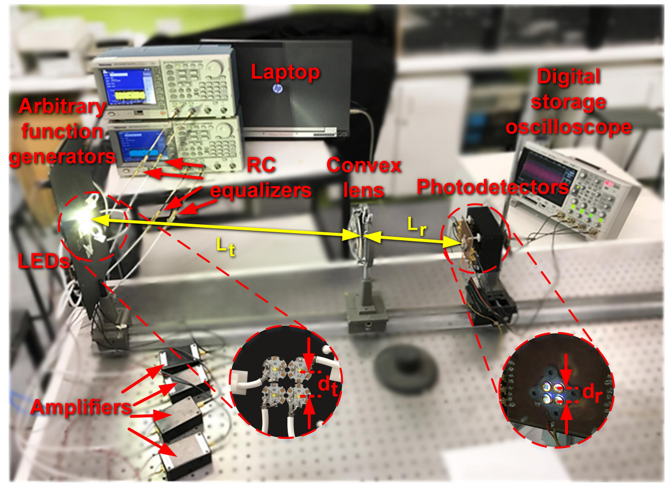


Fig. 3. Experimental setup. The convex lens is used to map the LEDs to the PDs. Spacing between Rxs and Txs d_r and d_t, respectively

where B_{sig} is given by:

$$B_{sig} = \frac{1}{T_s}(1+\beta)m. \tag{7}$$

Considering all subcarriers, the *m*-CAP output signal is given as [32]:

$$S(t) = \sqrt{2}(S_{I,f}(t) - S_{Q,f}(t)). \tag{8}$$

The *m*-CAP signal is then applied to identical arbitrary function generators (AFGs) using LabVIEW to generate the signals for the 4 spatial channels. The outputs $s(t)$ of the four channels are sampled at 2 GS/s with a vertical resolution of 8-bit and is then passed through four independent resistor-capacitor based pre-equalizers (RC EQ) in order to extend B_{LED} from $\sim$2 MHz to 3.8 MHz, as shown in Fig. 2.

The four equalised signals $s_{eq}(t)$ are then amplified (with a gain of 5 dB) prior to intensity modulation of LEDs (4 cool white 565K LUXEON Rebel LEDs) using a bias tee. Fig. 3 illustrates the experimental setup of the proposed system.

Note, the distances between the transmitter (Tx) and the lens L_t and between the lens and the optical receiver (Rx) L_r, respectively are given as:

$$L_t = f_l\frac{(\alpha-1)}{\alpha}, \tag{9}$$

$$L_r = f_l(1-\alpha), \tag{10}$$

where f_l is the lens focal length with a magnification factor given by [34]:

$$\alpha = -\frac{d_r}{d_t} = -\frac{L_r}{L_t}. \tag{11}$$

where d_r and d_t are the spacing between Rxs and Txs, respectively.

Using the parameters given in Table I, the total link span $L = L_t + L_r = 4.5f_l$. For f_l of 0.25, 0.4 and 1 m, L are 1.125 m, 1.8 m and 4.5 m, respectively, which are representative dimensions for a typical indoor environment.

As shown in Fig. 3, at the Rx, a convex lens with a focal length f_l is used to focus the received optical beams onto photodetectors (PDs) (silicon PD OSD15-5T). The received signal is given by:

$$y(t) = \mathcal{R}s_{eq}(t) \otimes h(t) + n(t), \tag{12}$$

where $h(t)$ is the channel impulse response, $\mathcal{R}$ represents the PD's responsivity, and $n(t)$ is the additive white Gaussian noise (AWGN). Note, the dominant noise sources are the ambient induced shots noise and the thermal noise, which are defined, respectively in terms of variances, as given in [28]:

$$\sigma_n^2 = 2qI_{bq}B_{Rx} + 4kTB_{Rx}, \tag{13}$$

where I_{bg} is the background light induced current, q is the electron charge, T is the absolute temperature, k is the Boltzmann's constant, and B_{Rx} is the receiver bandwidth.

TABLE I
SETUP AND PD PARAMETERS

Parameter	Value
Data-NRZ pseudorandom binary sequences length	$2^{17}-1$
Amplifier gain	5 dB
Convex lens • Focal length f_l • Diameter d	 0.25 m, 0.40 m and 1 m 110 mm, 105 mm and 50 mm
LED bandwidth B_{LED}	1.9 MHz
Photodetector - silicon OSD15-5T • Active area • Responsivity $\mathcal{R}$ • Bandwidth B	 15 mm^2 0.21 A/W 29 MHz
Spacing between Txs d_t	0.01 m
Spacing between Rxs d_r	0.02 m
Signal bandwidth B_{sig}	5, 10, 15, and 20 MHz
Transmission span L	$\sim$1 m, 1.8 m and 4.5 m

A typical value for the thermal noise σ^2_{ther} = 17.5348×10^{-16} A^2 [35]. For a typical indoor environment the standard illumination level is between 300 - 500 lux at 0.8 m height from the floor level, which is equivalent to a power requirement of $> 2.25 \times 10^{-4}$ W [35]. This represents the shot noise variance of at least $\sim 2 \times 10^{-16}$ A^2, which is the same as σ^2_{ther}.

Following optoelectronic conversion and amplification using transimpedance amplifiers (TIAs), the regenerated electrical signals $s_r(t)$ are captured using a digital storage oscilloscope (DSO-X3034A) with a sampling rate of 4 GS/s for off-line processing as outlined in the following paragraphs. $s_r(t)$ is first passed through a 4$^{\text{th}}$ order Butterworth LPF with a cut-off frequency f_{cut} = B_{sig} [26] in order to reject the out-of-band noise. The filtered m-CAP signal is then resampled to match the transmitted sampling frequency prior to matched filtering, the output of which is given by:

$$y_{m-I}(t) = y_s(t) \otimes f_{I,i}(-t), \quad (14)$$

$$y_{m-Q}(t) = y_s(t) \otimes f_{Q,i}(-t), \quad (15)$$

where $y_s(t)$ is the resampled signal. $f_I(-t)$ and $f_Q(-t)$ are the impulse responses of the matched filters.

Following M-QAM demodulation and parallel-to-serial conversion (P/S) the estimated data symbols $d'_d(t)$ are recovered. As reported in the literature, a binary phase shift keying based signal is first transmitted on each subcarrier to measure the error vector magnitude (EVM), prior to SNR estimation following the procedures adopted from [36], [37]:

$$\mathrm{SNR}_i = 20\log_{10}\left(\frac{\mathrm{EVM}_{\mathrm{RMS}_i}(\%)}{100}\right). \quad (16)$$

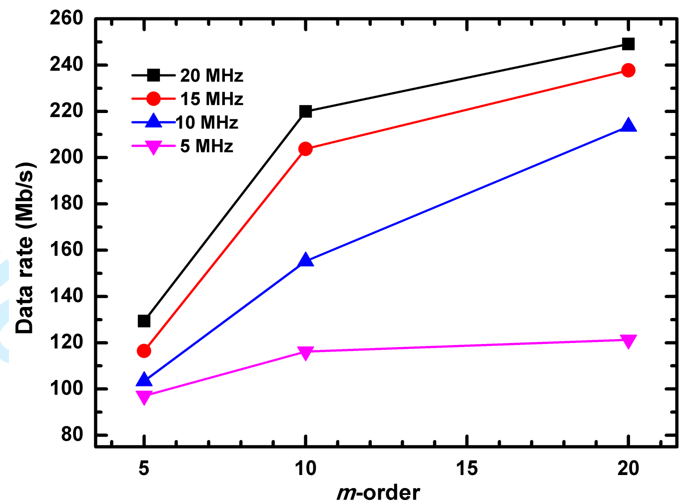
Next, a subcarrier-specific value of b is selected based on the estimated SNR. The proposed system performance is then assessed in terms of (*i*) R_b as a function of m; and (*ii*) the assigned b versus the order of m. Here, we have adopted the target BER of 10^{-3} allowing a margin for the 7% forward error correction limit (FEC), which has a BER limit of 3.8×10^{-3} [38]. The SNR$_{\text{threshold}}$ for the targeted BER can be found as the literature [39]. Therefore, at the target BER we defined a set of values for SNR$_{\text{threshold}}$ of $\{6.8, 9.8, 16.6, 22.6, 28.5\}$ dB for $b = \{1, 2, 4, 6, 8\}$, respectively. Note, $R_{\text{net}} = \sum R_b$ for all four channels after removal of the 7% FEC overhead. Following the above procedure, the m-CAP signals are transmitted and the BER is determined by comparing $d'_d(t)$ and $d_d(t)$.

III. Results and Discussion

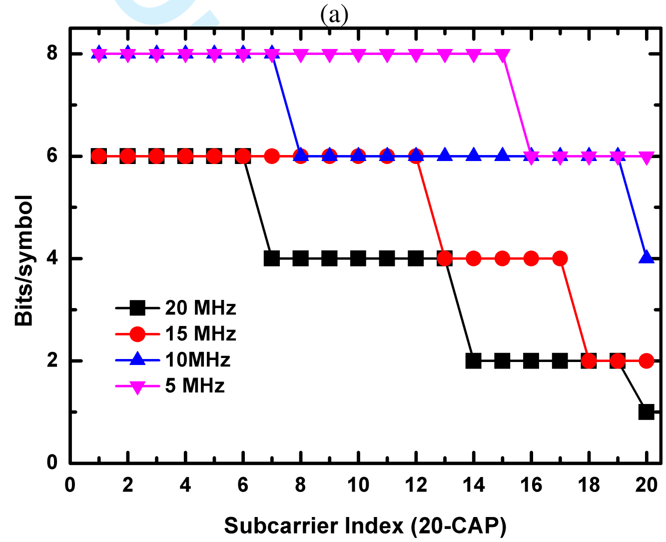
In this section, we outline experimental evaluation of a 4×4 optical imaging MIMO-VLC utilising the m-CAP scheme by investigating the impact of using (*i*) different B_{sig}, while maintaining a fixed L of $\sim$1 m; (*ii*) a range of subcarriers m up to 20 and B_{sig} of 5, 10, 15 and 20 MHz for L of $\sim$1 m at the adopted BER target of 10^{-3}; and (*iii*) a range of L = $\sim$1, 1.8, and 4.5 m and $m \in \{20, 15, 10, 5\}$ when B_{sig} is 10 MHz.

A. Signal bandwidth B_{sig}

Fig. 4(a) illustrates R_{net} as a function of m for a range of B_{sig}, for L of $\sim$1 m and a BER of 10^{-3}. Note, 20-CAP displays the highest R_{net} of $\sim$249 Mb/s (the highest reported in this work) at BER of 3.2×10^{-3} for B_{sig} = 20 MHz, which reduces to $\sim$121 Mb/s for B_{sig} = 5 MHz. At B_{sig} of 5 MHz, the increase in R_{net} as a function of m is relatively small since all the subcarriers have a similar bandwidth per subcarrier B_{SC}. For B_{sig} of 20 MHz, the drop in R_{net} is due to reduced B_{SC} for a given m and B_{sig}. For example, for m = 20 and B_{sig} = 20 MHz, B_{SC} is 1 MHz, whereas, for m = 20 and B_{sig} = 5 MHz, B_{SC} reduces to 0.25 MHz. For B_{sig} of 15 and 10 MHz, R_{net} reaches 237.75 Mb/s and 213.49 Mb/s, respectively. Moreover, for higher values of B_{sig} (e.g., 20 MHz and m = 20) the out of band subcarriers can still be loaded with b = 4 starting from the 7$^{\text{th}}$ subcarrier as shown in Fig. 4(b), which depicts the assigned b values for 20-CAP and for



(a)



(b)

Fig. 4. (a) The net data rate R_{net} against m for a range of B_{sig}, and a BER of 10^{-3}, and (b) assigned bits/symbol for 20-CAP for a range of B_{sig} (20, 15, 10 and 5 MHz). Note, in (b) the highest value of b = 8 is assigned for the first 7$^{\text{th}}$ and 15$^{\text{th}}$ subcarriers with B_{sig} of 10 and 5 MHz, respectively, since they are within B_{LED}

a range of B_{sig} {20, 15, 10 and 5 MHz}.

Using a range of B_{sig} (i.e., 20, 15, 10 and 5 MHz) for m = 20, we observe η_{net} of 12.45, 15.85, 21.34, 24.26 b/s/Hz, respectively. This is interesting as it shows that using higher B_{sig} leads to less utilisation of the spectrum but higher R_{net}. Moreover, for m = 20 and B_{sig} of 5, 10 and 15 MHz we observe an improvement in η_{net} of ~95%, ~72% and ~27% compared to m = 20 and B_{sig} of 20 MHz. Hence, there is a trade-off between the achievable R_{net} and spectral utilisation when considering different values of B_{sig} for the same m-CAP system. This can be attributed to the fact B_{SC} increases with B_{sig} since the subcarriers are divided equally, which leads to a reduced SNR of each individual subcarrier positioned outside the pass-band region (i.e., the roll-off region of the LPF response of the LED with 20 dB/decade). Therefore, they are assigned with a low b value (see Fig. 4(b)). However, due to the increase in B_{sig} the R_{net} is boosted as it increases proportionally with B_{sig} (i.e., $\frac{1}{T_S}(1+\beta)m$). Moreover, $R_{net} = \sum_{i=1}^{m} b_i \frac{1}{T_s}$ for all four channels in addition to removal of the 7% FEC overhead. In contrast, for a given m decreasing B_{sig} leads to a higher SNR per subcarrier, which is due to lower B_{SC} that ensures higher b values (Fig. 4(b)). Therefore, achieving higher spectral utilisation for a given B_{sig}.

Figs. 5(a) and (b) depict the measured frequency spectra of the 20-CAP system for B_{sig} of 20 MHz and the corresponding constellation diagrams of the 1st (black), 7th (red), and 14th (green) subcarriers, respectively. From Fig. 5(a) the 1st six subcarriers occupy a bandwidth of 6 MHz. Note, $\mathrm{EVM_{RMS}}$ values for the three constellations shown are 5.11%, 7.31% and 15.73%, respectively. Fig. 6 then illustrates the spectral efficiency as a function of m for the range of B_{sig}.Note, from Fig. 4 (a) for m = 10 and B_{sig} of 20, 15, 10 and 5 MHz, the R_{net} that can be supported are 219.96, 203.79, 155.26 and 116.25 Mb/s, respectively. This corresponds to η_{net} of 11, 13.58, 15.52 and 23.25 b/s/Hz for the listed B_{sig}. Moreover, improved spectral utilisation of ~130%, ~54% and 34.45% for B_{sig} of 5, 10 and 15 MHz, respectively, is obtained in contrast to B_{sig} of 20 MHz. For m = 5 and B_{sig} of 20, 15, 10 and 5 MHz, η_{net} values are reduced to 6.46, 7.76, 10.35 and 19.4 b/s/Hz, respectively. Alternatively, for B_{sig} of 5, 10, and 15 MHz and m of 5 we observe corresponding improvement in the spectrum utilisation of 203% ~60% and ~74% compared to B_{sig} of 20 MHz.

At this stage, it is worth mentioning that by combining both space multiplexing (MIMO) and frequency multiplexing (m-CAP) schemes, the overall system transmission speed is improved by ~8 times from 31.53 Mb/s to ~249 Mb/s in contrast to [26]. To the best of authors' knowledge this is the only experimental work on implementation of m-CAP (up to 10) in the VLC domain reported in the literature. We have kept L to be ~1 m. However, in this work B_{sig} is increased by ~3 times, which is higher than the 6.5 MHz adopted in [26]. It should be noted that, the aim of this paper is not to achieve the highest value in R_{net}, but to demonstrate the possible gain that can be achieved by combining space and frequency multiplexing schemes.

Moreover, in order to increase R_{net} we have doubled the m-order to 20, but at the cost of increased computational complexity of the overall system, since for every increment in m, 2 more pulse shaping FIR filters (i.e., one each for I and Q) are required at both the Tx and the Rx. Note, increasing m from 10 to 20 will lead to the increase of the number of convolution operations from 460 to 1840 at the Tx. For MIMO-VLC m-CAP with m = 10, R_{net} is increased to 213.49 Mb/s compared to 31.53 Mb/s [26] (i.e., a factor of 6.77) over the same L of 1 m. Increasing B_{sig} to 10 MHz results in η_{net} being increased by a factor of 4.4 as the obtained η_{net} is 21.34 b/s/Hz

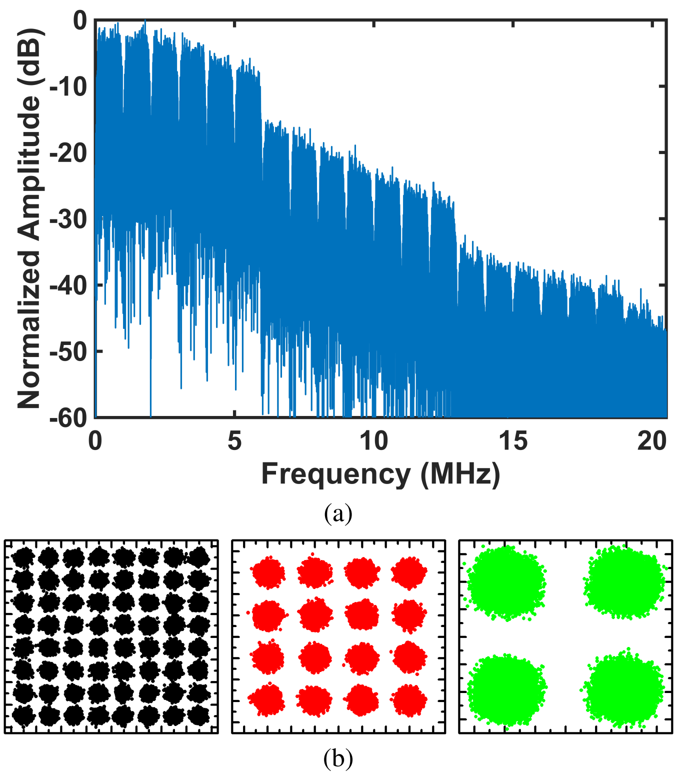


Fig. 5. 20-CAP with B_{sig} of 20 MHz; (a) the measured frequency spectrum, and (b) constellation diagrams for the 1st, 7th and 14th subcarriers

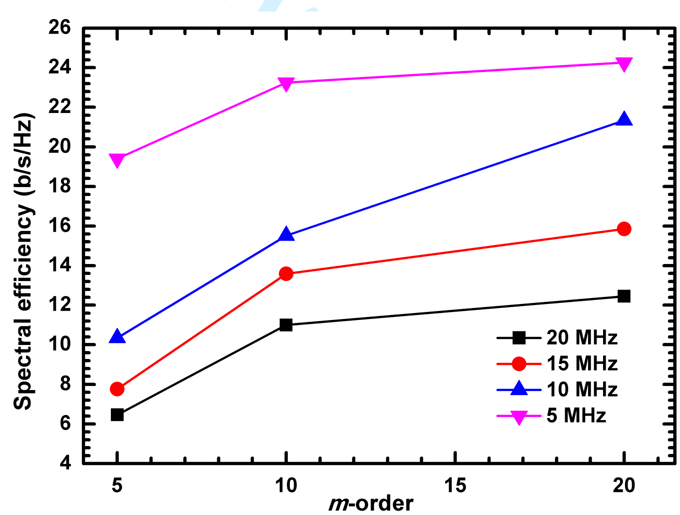


Fig. 6. The spectral efficiency against m for a range of B_{sig} {20, 15, 10 and 5 MHz}

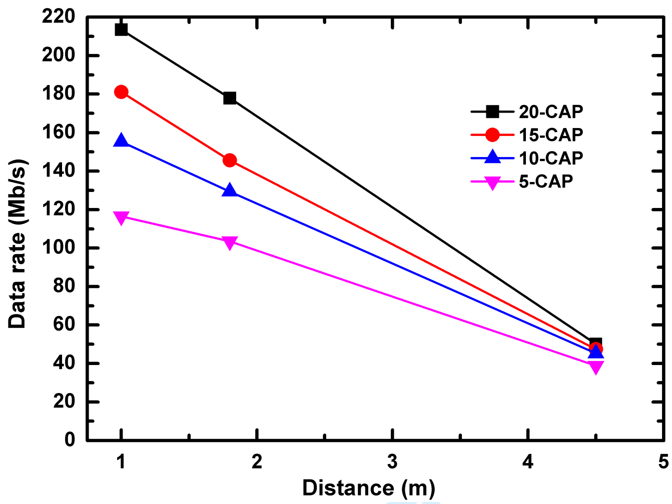


Fig. 7. The R_{net} as a function of the transmission distance for B_{sig} of 10 MHz and a range of m

compared to 4.85 b/s/Hz in [26], which is considerable in VLC.

B. Transmission span

Fig. 7 illustrates R_{net} as a function of L for a range of $m \in \{20, 15, 10, 5\}$ and B_{sig} of 10 MHz. As expected, with the increasing L from 1 to 4.5 m has reduced R_{net} from 213 Mb/s to approximately 50 Mb/s for 20-CAP. Therefore for 20-CAP this results in η_{net} of 21.35 b/s/Hz and 5 b/s/Hz for the BER values of 1.2×10^{-3} and 3.7×10^{-3}, respectively. For a range of $m \in \{20, 15, 10, 5\}$ and L of $\sim$1 m R_{net} values are 213.49, 181.14, 155.26 and 116.44 Mb/s, respectively; where the corresponding values of η_{net} are 21.35, 18.11, 15.52 and 11.64 b/s/Hz. Finally, Fig. 8 outlines the achievable BER performance as a function of L for a range of m. As expected, the BER increases with L and m, reaching 3.7×10^{-3}, which is still below the 7% FEC limit of 3.8×10^{-3} for m = 20 and L of 4.5 m.

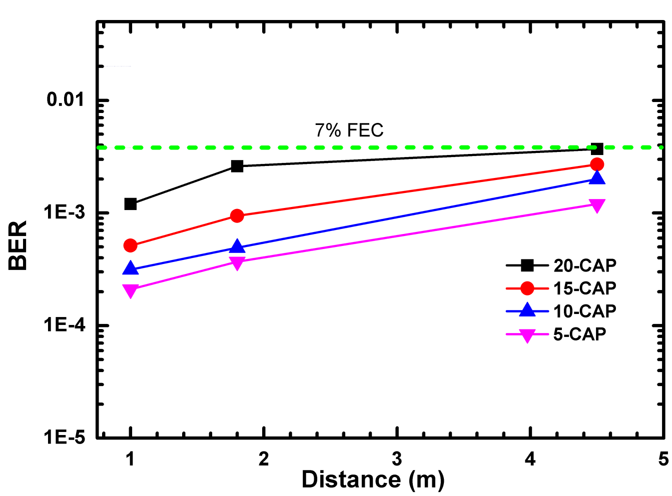


Fig. 8. The measured BER as a function of the distances for different values of m; also included is the 7% FEC limit (dashed green)

By extending L to 1.8 m, R_{net} is dropped to 177.9, 145.56, 129.39 and 103.5 Mb/s (see Fig. 7); which corresponds to a reduction in η_{net} of 17.79, 14.55, 12.94 and 10.35 b/s/Hz. The highest BER achieved for the L of 1.8 m is 2.6×10^{-3} at m = 20 (see Fig. 8); following the same upward trend as in m and L.

For the highest L of 4.5 m, R_{net} is reduced (refer to Fig. 7) compared to the shorter distances, with R_{net} of 50.13, 47.43, 45.28 and 38.81 Mb/s at $m \in \{20, 15, 10, 5\}$, respectively. Similar values of R_{net} are observed for lower values of b for each subcarrier. In other words, the bit loading algorithm fails to introduce any improvement in the overall R_{net} as a result of all subcarriers having low SNR values (i.e., $\ll$ 16.6 dB; which is the $\mathrm{SNR}_{\mathrm{threshold}}$ at b = 4). Also, a reduction in the spectral efficiency was observed and for the first time in this work, it fell below 5 b/s/Hz.

IV. CONCLUSION

In this paper, we have reported for the first time, an investigation of the performance of a 4 $\times$ 4 m-CAP imaging MIMO (i.e., frequency and space multiplexing) VLC link considering a range of system parameters. By combining the two schemes, we experimentally demonstrated an improvement in the transmission speed reaching a maximum value of $\sim$249 Mb/s. We showed that there is a trade-off between R_{net} and η_{net}, since for a given values of m and L increasing B_{sig} substantially improved R_{net}; however, interestingly η_{net} was reduced. Moreover, the highest R_{net} of 249.07 Mb/s was achieved at a BER of 3.2×10^{-3} for B_{sig} of 20 MHz and for m = 20 over a transmission span of $\sim$1 m. On the other hand, for lower B_{sig} of 5 MHz and for m = 20 the achieved R_{net} was reduced to 121.21 Mb/s, but spectrum utilisation was almost 95% better in contrast to B_{sig} of 20 MHz. We also observed that for a given value of m and B_{sig}, increasing the distance resulted in reduced R_{net}. The overall R_{net} (η_{net}) for all proposed system (i.e., four channels) were 213.49 Mb/s (21.35 b/s/Hz), 177.9 Mb/s (17.79 b/s/Hz), and 50.13 Mb/s (5.01 b/s/Hz) for m = 20 and B_{sig} of 10 MHz over a range of L (i.e., $\sim$1 m, 1.8 m and 4.5 m), respectively, at BER values of 1.2×10^{-3}, 2.6×10^{-3}, and 3.7×10^{-3}.

Moreover, we showed that for the highest L of 4.5 m increasing m did not introduce any significant improvement in the system performance, but increased the computational complexity by the way of increased number of the pulse shaping FIR filters. In our future work, we will experimentally compare the m-CAP and OFDM schemes performance.

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