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Performance of OFDM with enhanced Subcarrier Index Modulation using QAM

Nimitha N*[1], Sesha Vidhya[2], Parvathy G[3], Sathish Kumar[4],
R.M.K College of Engineering and Technology,Puduvoyal, Gummidipoondi Taluk,Thiruvallur District,India
[1,3, 4] Assistant Professor, Department of Electronics and Communication Engineering, Anna University
[2]Associate Professor, Department of Electronics and Communication Engineering, Anna University

ABSTRACT:- Recently, orthogonal frequency division multiplexing (OFDM) with index modulation (OFDM-IM) using QAM was proposed. By selecting a fixed number of subcarriers as active subcarriers to carry constellation symbols, the indices of these active subcarriers may carry additional bits of information. In this paper, we propose two generalization schemes of OFDM-IM, named as OFDM with generalized index modulation 1 (OFDMGIM1) and OFDM-GIM2, respectively. In OFDM-GIM1, the number of active subcarriers in an OFDM sub block is no longer fixed. Dependent on the input binary string, different number of active subcarriers are assigned to carry constellation symbols In OFDM-GIM2, independent index modulation is performed on the in-phase and quadrature component per subcarrier. Through such ways, a higher spectral efficiency than that of OFDM-IM may be achieved. Since both generalization schemes proposed suffer from BER performance loss in low SNR region, an interleaving technique is proposed to tackle this problem. Finally, noting that the two generalization schemes are compatible with each other, the combination of these two schemes, named as OFDM-GIM3, has also been investigated. Computer simulation results clearly show our proposed scheme's superiority in both spectral efficiency and BER performance compared to existing works.
Index Terms-Bit Error Rate(BER), generation, index modulation,Interleaver, LLR technique, orthogonal division multiplexing(OFDM), spectral efficiency

I. INTRODUCTION

Orthogonal Frequency Division Multiplex or OFDM is a modulation format that is finding increasing levels of use in today's radio communications scene. OFDM has been adopted in the Wi-Fi arena where the 802.11a standard uses it to provide data rates up to 54 Mbps in the 5 GHz ISM (Industrial, Scientific and Medical) band. In addition to this the recently ratified 802.11g standard has it in the 2.4 GHz ISM band. In addition to this, it is being used for WiMAX and is also the format of choice for the next generation cellular radio communications systems including 3G LTE and UMB.If this was not enough it is also being used for digital terrestrial television transmissions as well as DAB digital radio. A new form of broadcasting called Digital Radio Mondiale for the long medium and short wave bands is being launched and this has also adopted COFDM. Then for the future it is being proposed as the modulation technique for fourth generation cell phone systems that are in their early stages of development and OFDM is also being used for many of the proposed mobile phone video systems. OFDM, orthogonal frequency division multiplex is a rather different format for modulation to that used for more traditional forms of transmission. It utilizes many carriers together to provide many advantages over simpler modulation formats. An OFDM signal consists of a number of closely spaced modulated carriers. When modulation of any form - voice, data, etc. is applied to a carrier, then sidebands spread out either side. It is necessary for a receiver to be able to receive the whole signal to be able to successfully demodulate the data. As a result when signals are transmitted close to one another they must be spaced so that the receiver can separate them using a filter and there must be a guard band between them. This is not the case with OFDM. Although the sidebands from each carrier overlap, they can still be received without the interference that might be expected because they are orthogonal to each another. This is achieved by having the carrier spacing equal to the reciprocal of the symbol period.

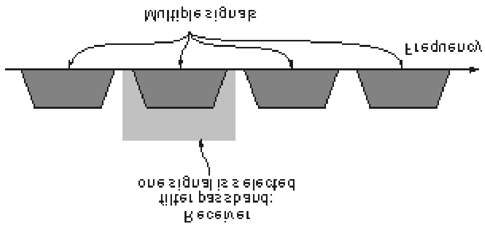


Fig 1. Traditional view of receiving signals carrying modulation

To see how OFDM works, it is necessary to look at the receiver. This acts as a bank of demodulators, translating each carrier down to DC. The resulting signal is integrated over the symbol period to regenerate the data from that carrier. The same demodulator also demodulates the other carriers. As the carrier spacing equal to the reciprocal of the symbol period means that they will have a whole number of cycles in the symbol period and their contribution will sum to zero - in other words there is no interference contribution.

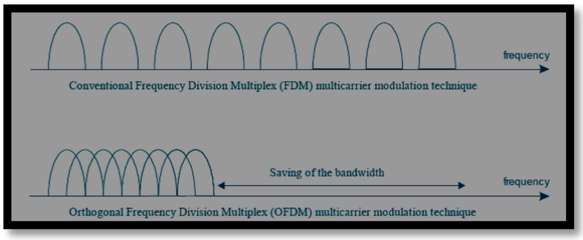


Fig 2. OFDM Spectrum

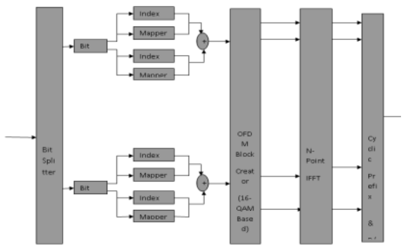


Fig. 3. Block diagram of the OFDM-IM transmitter in [1]

One requirement of the OFDM transmitting and receiving systems is that they mut be linear. Any non-linearity will cause interference between the carriers as a result of inter-modulation distortion. This will introduce unwanted signals that would cause interference and impair the orthogonality of the transmission. In terms of the equipment to be used the high peak to average ratio of multi-carrier systems such as OFDM requires the RF final amplifier on the output of the transmitter to be able to handle the peaks whilst the average power is much lower and this leads to inefficiency.

II. THE FIRST GENERALIZATION OF OFDM-IM

In the OFDM-IM scheme, the number of active subcarriers in each OFDM subblock is fixed to a value of K and the best performance in terms of BER and spectral efficiency is achieved when K = n=2 if QAM is used. The spectra efficiency is defined to be B=(N + L) in [1]. For the case of n = 8 and n = 32, the maximum spectral efficiency achieved are 1.1111 bits/s/Hz and 1.25 bits/s/Hz, respectively, and the BER performance outperforms that of classical OFDM for signal-to-noise ratio (SNR) higher than 30 dB. However, the number of active subcarriers in an OFDM subblock is not necessarily fixed. In the following, the OFDM-IM is generalized such that the number of active subcarriers in an OFDM subblock may be different. This generalization is named as OFDM with generalized IM1, denoted as OFDMGIM1.

A. Main Idea For an OFDM sub block, let n be the total number of subcarriers of the sub block. The number of subcarriers that are active to carry constellation symbols may vary dependent on the input signal to be transmitted. Let K be the set of all allowed numbers of active subcarriers, and R is defined as the size of set K. For example, if 1 or 3 out of n subcarriers are allowed to carry signal constellation symbols, K = f1; 3g and R is equal to 2. Thus, for an OFDM sub block, K is given by,

$$K = \{K1, K2, \ldots., KR\} \qquad (12)$$

In the case of Kr active subcarriers for Kr 2 K and r = 1,......,R, the indices of the selected Kr active subcarriers of the g-th subblock, denoted as I, is given by,

$$I_r^g = \{i_{r,1}^g, \ldots., i_{r,Kr}^g\} \qquad (13)$$

whereigr;k2 f1; : : : ; ngfor g = 1; : : : ;G, k = 1; : : : ;Kr and ig r;k1 6= ig r;k2 if k1 6= k2.Similarly, the signal constellation symbols at the output of the mapper to be put onto the subcarriers with indices in Ig r , denoted as Sg r , is given by

$$S_r^g = \{s_{r,1}^g, \dots \dots \dots \dots, s_{r,Kr}^g\} \quad (14)$$

In the OFDM-IM scheme proposed in [1], once n is given, K is fixed and only has a single element for all the subblocks. As a result, the OFDM-IM scheme is a special case of our OFDM-GIM1 scheme. For a certain Kr 2 K, the total number of bits that can be transmitted by an OFDM sub block is given by

$$B_r^g = \lfloor \log_2 (M^{Kr} C_n^{Kr} \rfloor \quad (15)$$

and the total number of bits that can be transmitted by all Kr 2 K of an OFDM subblock is given by

$$\sum_{Kr \in K} B_r^g = \lfloor \log_2 (\sum_{Kr \in K} M^{Kr} C_n^{Kr}) \rfloor \quad (16)$$

Considering an extreme case where K = {0; 1; : : : ; n} we have,

$$\sum_{r=1}^{n+1} B_r^g = \log_2 (M+1)^n \quad (17)$$

Clearly, log2(M +1)n is much larger than Bg r , the number of bits that can be transmitted by a fixed Kr. This extreme case indicates that by allowing multiple values of K, we can obtain much more possible ways of selecting active subcarriers for a given sub block size n. Hence, compared to the OFDM-IM scheme, more information bits per sub block can be transmitted in our OFDM-GIM1 scheme. For example, the spectral efficiency of our OFDM-GIM1 scheme will be up to 20% higher than that of OFDM-IM when n = 8. However, compared to the implementation of OFDM-IM, several changes are necessary to ensure the successful implementation of our proposed OFDM-GIM1 scheme.1) The normalization factors in [1] needs to be adjusted. To ensure E_ xHx equal to N, E _ sgsgH is set to np. And the normalization factor of IFFT is changed to N and accordingly the normalization factor of FFT is changed to 1= p N. 2) The index modulation block in Fig. 1 works for K containing only a single element. In this case, the p bit input binary string is split to fixed p1 bits and p2 bits for a given K value, regardless of the information represented by the string. Generalizing this block is necessary so that p bit input string with different values may be split to different p1 bits and p2 bits, and thus to use different number of active subcarriers to carry M-ary constellation symbols for a given set K with multiple elements. In other words, for OFDM-GIM1, p remainsconstant in each transmission subblock while p1 and p2 may vary according to different incoming input strings.

3) The LLR detector at the receiver to detect the active subcarriers and the M-ary constellation symbols carried by the active subcarriers are upgraded to adapt to a given
set K with multiple elements.

As we can see, compared to OFDM-IM, 1) is a minor change and 2) and 3) are major changes. The details of these two major changes are addressed in the following 2 subsections.

A) Generalized Index Modulation Block

In this subsection, we first take n = 8;K = f1; 3; 5g and M = 2 (i.e., QAM) as an example to demonstrate how the generalized index modulation block works. The elements of the set K are assumed to be in strict ascending order. These parameters mean that for the g-thsubblock consisting of 8 subcarriers, either a subcarrier, 3 subcarriers or 5 subcarriers out of 8 subcarriers may carry QAM symbols. Hence, according to (16), 11 bits, i.e., p = 11, can be transmitted by the g-thsubblock. To map the 11 bits to the subcarrier indices and QAM symbols, our method starts with K1 = 1, i.e., a subcarrier out of 8 subcarriers is chosen and a QAM symbol is put onto the selected subcarrier. From MK1CK1 n = 21C1 8 = 16, it is known that in such a way 16 combinations out of 2p = 211 = 2048 combinations can be expressed. The 16 combinations are listed in Table I. Table I consists of three columns. The first column (in italic style) contains the sequence numbers of the combinations, Zp, which are also the decimal expressions of the binary strings to be transmitted by the subblock. The second column (in bold style) and the third column (underlined) together are the binary strings to be transmitted (arranged in ascending order). Since only one subcarrier is active when K1 = 1, the number of the information bit which can be carried by the active subcarrier in QAM symbol, denoted as p2, is one, i.e., p2 = 1 in this case. Taking the p2 least significant bit (in column three) to be the bit carried by the active subcarrier, the remaining p1 more significant bits (in column two) are mapped to determine the subcarrier index that is active. The combinatorial method introduced in [1], [19], [20] may be used for the mapping. The combinatorial method suggests that the range of input decimal numbers to be mapped should be a contiguous integer set starting from zero. For OFDM-IM, since there is only one element in K, i.e., all input decimal numbers are mapped based on a fixed number of active sub carriers, the above requirement is automatically satisfied. However, for our proposed OFDM-GIM1, only when the input numbers are mapped based on the first element in K, i.e., mapped to K1 active subcarriers, the requirement for the range of input numbers is satisfied for sure. The range of input numbers mapped based on the other elements, i.e., Kr 2 K; r = 2; : : : ;R, is not necessarily starting from zero. As a result, an offset is introduced for each Kr to ensure that the range of the input decimal numbers satisfies the requirement of the combinatorial method.

Let Zp1 represent the decimal number of the p1 bit binary string, and denote Zr p1 as the decimal number of the first combination of the p1 bits carried by Kr active subcarriers (see Z1 p1 in Table I). This Zp1 is the offset to make the range of input numbers satisfy the requirement of the combinatorial method. Offsetting Zp1 by a value Zrp1 , the resultant value, denoted as Z0 p1 , is Z0 p1 = Zp1 □Zr p1 . For instance, for Zp= 2, we have Zp1 = 1;Z1p1 = 0 and Z0 p1 = Zp1 □Z1 p1 = 1. This Z0 p1 , together with n = 8, K1 = 1 are fed to the combinatorial method proposed in [1], [19], [20]. The output of the combinatorial method for this example is Ig1 = f2g, indicating that the second subcarrier in this subblock is the selected active subcarrier. Meanwhile, the p2 bit underlined, i.e., 0, is mapped to QAM symbol sg1 = [□1] and to be put onto the second subcarrier. Once the mapping for K1 = 1 is done, our method then proceeds to K2 = 3, where 3 subcarriers out of 8 subcarriers are chosen and 3 QAM symbols are put onto those selected subcarriers. The total number of combinations that K2 = 3 can represent is given by:

$$\boldsymbol{M^{K_2} C_n^{K_2} = 2^3 C_8^3 = 448} \qquad (18)$$

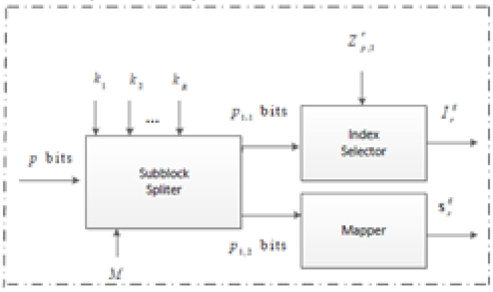


Fig 4: The generalized index modulation block of the g-th subblock for OFDM-GIM1 using QAM

The 448 combinations are given in Table II. The procedure of the combination mappings for K2 = 3 is similar to that of K1 = 1. For example when Zpis 462, p2 = 3, p1 = 8, Z2 p1 = 2 and the offset decimal number Z0 p1, given by Zp1 -Z2 p1 , thus is 57-2 = 55. This number, together with n = 8, K2 = 3 are fed to the combinatorial method. The output of
the combinatorial method will be Ig2 = f6; 7; 8g, i.e., the sixth, seventh and eighth subcarriers are selected in this subblock as the active subcarriers. Meanwhile, the p2 bits in underlined type, i.e., 110, are mapped to QAM symbols sg2 = [1; 1;1] and to be put onto those selected subcarriers respectively. The procedure of the combination mappings for K3 = 5 continues. Theoretically, the total number of combinations that K3 = 5 can represent is given by:

$$\mathbf{M^{K_3} C_n^{K_3} = 2^5 C_8^5 = 1792} \qquad \mathbf{(19)}$$

As 464 out of 2048 combinations have already been covered by K1 and K2 active subcarriers, only the remaining 1584 combinations are to be covered by using K3 = 5: By noting this, the procedure of the combination mappings for K3 = 5 is the same as that of K1 = 1 and K2 = 3. Through this example, it can be seen that the proposed index modulation block has one more level of operation than the original one. In the original technique, the number of active subcarriers is fixed, such that the splitting of p bits to p1 and p2 bits is also fixed. However, in the proposed technique, the p bit signal is split into different p1 and p2 bits for different input binary strings for a given set K. For a certain Zp, the number of bits p2 entering the mapper is given by,

$$\boldsymbol{p_2} = \begin{cases} \boldsymbol{K_1 \ \log_2 M, \quad Z_p \epsilon [0, M^{K1} C_n^{K1} - 1]} \\ \boldsymbol{K_2 \ \log_2 M, \quad Z_p \epsilon [M^{K1} C_n^{K1}, M^{K2} C_n^{K2} - 1]} \\ . \\ . \\ \boldsymbol{K_{R2} \ \log_2 M, \quad Z_p \epsilon [M^{KR-1} C_n^{KR-1}, M^{KR} C_n^{KR} - 1]} \end{cases} \qquad (20)$$

In the example mentioned above, we have

$$\boldsymbol{p_2} = \begin{cases} \boldsymbol{1, \quad Z_p \in [0, 15]} \\ \boldsymbol{3, \quad Z_p \in [16, 463]} \\ \boldsymbol{5, \quad Z_p \in [16, 463]} \end{cases} \qquad (21)$$

Thus, the index modulation block in Fig. 1 is modified to the generalized structure in Fig. 2, where the subblock splitter takes the p bit signal, the set K and M as input to determine the value of p2 for this particular p bit input (with decimal value Zp) according to (20). Once p2 is determined, the number of bits entering the index selector is given byp1 = p-p2. Thereafter, the index selector takes the p1 most significant bits of the input binary string and generates theindices of p2 active subcarriers Igr using the combinatorialmethod with the offset Zrp1 considered, where r satisfiesp2 = Kr and r 2 f1; : : : ;Rg. Meanwhile, the p2 leastsignificant bits are sent to the mapper and are mapped to M-ary constellation symbols Sgr . The constellation symbols Sgrare then put onto the active subcarriers with indices in Igrrespectively.

B . Upgraded LLR detector

The ML detector is optimum in the detection of received symbols in OFDM-IM, as the ML detector considers all possible subblock realizations by searching for all possible subcarrier index combinations. However, in [1], ML detector may only be applied to the cases where $\mathbf{M^K C_n^K}$is small because large $\mathbf{M^K C_n^K}$values produce a large number of possible subcarrier index combinations. Similarly, our proposed scheme may use ML detector when the number of possible subcarrier index combinations is small. However, in general, our proposed scheme has many more combinations than OFDM-IM, mPaking ML detector generally impractical in most cases where $\sum_{r=1}^{R} \mathbf{M^{K_r} C_n^{K_r}}$is large. According to [1], LLR detector is a practical choice to trade off between the detection precision and detection complexity. In this subsection, the upgraded LLR detector for OFDM-GIM1 is proposed.

In the original OFDM-IM [1], for the g-thsubblock, there is only a single element in the set K, i.e., K = fK1g. In the receiver, once the K1 indices of the subcarriers which have the maximum K1 LLR values out of the n LLR values, computed according to (11), are obtained, the demodulation of M-ary constellation symbols is straightforward from those chosen subcarriers [1]. Different from OFDM-IM, our generalized scheme has a flexible Kr 2 K for p bit input signal with different values in an OFDM subblock. Though the set K is known by the receiver in advance, for each received symbol, the receiver actually does not know what Kr is. To detect the information, every possible Kr must be considered.

The detection procedure is done subblock by subblock. Takethe g-thsubblock as an example. The detection procedurestarts by calculating the LLR values of all the n subcarriersin the g-thsubblock. According to (11), for every Kr 2 K,we have

$$\lambda_r^g(\xi) = \ln\left(\sum_{m=1}^{M} exp\left(-\frac{1}{N_{0,F}}\left|Y^g(\xi) - \sqrt{\frac{n}{K_r}}H^g(\xi)s_m\right|^2\right)\right) + |Y^g(\xi)|^2 + \ln(K_r) - \ln(n-K_r)$$

(22)

$$Y^g(\xi) = Y(n(g-1)+\xi) \qquad (23)$$

$$H^g(\xi) = H(n(g-1)+\xi) \qquad (24)$$

$$\xi = 1, \ldots\ldots, n. \qquad (25)$$

and

$$s_m \epsilon \boldsymbol{S} \qquad (26)$$

Here, Y g(**ξ**) is the _-th received signal in the g-thsubblock,andHg (**ξ**) is the **ξ** -th channel fading coefficient in the g-thsubblock. In (22), a factor $\sqrt{\frac{n}{K_r}}$is introduced to normalize the received signal according to the assumed transmitted symbol. In OFDM-IM, no such factor is used in the LLR detector because K is fixed in OFDM-IM, and the normalization by a factor of $\sqrt{\frac{n}{K_{tot}}} = \sqrt{\frac{n}{K}}$has been done collectively before the LLR detector. For QAM modulation, by using the Jacobian logarithm [21] to prevent numerical overflow, (22) can be further simplifiedto

$$\lambda_r^g(\xi) = \max(a,b) + \ln\left(1 + \exp(-|b-a|)\right) + \frac{|Y^g(\xi)|^2}{N_{0,F}} + \ln(K_r) - \ln(n-K_r) \qquad (27)$$

Where

$$a = -\left|Y^g(\xi) - \sqrt{\frac{n}{K_r}}H^g(\xi)\right|^2 / N_{0,F} \qquad (28)$$

$$b = -\left|Y^g(\xi) - \sqrt{\frac{n}{K_r}}H^g(\xi)\right|^2 / N_{0,F} \qquad (29)$$

Based on the obtained Ig r and Sg r for all r, Ig^r and sg^r are regarded as the set of active subcarrier indices and the M- ary constellation symbols carried by these subcarriers if the distance between the assumed transmitted signal (with channel conditions considered) and the received signal is minimum among all r, i.e.,

$$\left(I_r^g, s_r^g\right) = \arg\min_{r\epsilon[1,\ldots,R]}\left(\sum_{\xi=1}^{n}|Y^g(\xi)|^2 + \sum_{k=1}^{K_r}\left|Y^g(I_r^g(k)) - H^g(I_r^g(k)s_r^g(k)\right|^2\right) \qquad (30)$$

Thereafter, the obtained Ig^r and sg^r are passed to the index demodulation block at the receiver which performs the opposite action of the index modulation block in Fig. 2, to provide an estimate of the p bit input binary string.

III. THE SECOND GENERALIZATION SCHEME OF OFDM-IM:

OFDM-IM successively increases the transmitting spectral efficiency and meanwhile improves the BER performance for signals with high SNR under QAM symbols. However, OFDM-IM becomes ineffective when higher constellation symbols other than QAM are implemented. For example, henM-ary constellation symbols are implemented, for a certain n and K, the total number of bit combinations that OFDM-IM can represent is given by MKCK n G. However, to achieve the same spectral efficiency as classical OFDM using QPSK, the total number of combinations required is MnG. noting that

$$\frac{M^n G}{M^k C_n^K G} = \frac{M^{n-K}}{C_n^K} > 1 \qquad (31)$$

for most choices of n and K when M _ 4, it is important to find schemes to remedy this shortcoming of OFDM-IM. Compared to classical OFDM using QAM, classical OFDM using QPSK doubles the spectral efficiency with no BER performance loss. The generalization technique proposed in Section III mitigates the problem by further improving the spectral efficiency with marginal BER performance loss. However, the proposed OFDM-GIM1 cannot fundamentally solve the problem for QPSK symbols. The second generalization of OFDM-IM, aiming at QPSK constellation symbols, is introduced in this section. This scheme is based on the original OFDM-IM and named as OFDM-GIM2.Main Idea: In wireless communication, anM-ary complex constellation symbol (for M _ 4) consists of an in-phase component and a quadrature component. In [1], the in-phase and quadrature components are regarded as inseparable and index modulation is applied coherently to the complex constellation symbol as a whole. In other words, in the original OFDM-IM, if a subcarrier is inactive, both the in-phase and quadrature components carried are 0, and if a subcarrier is active, both the in-phase and quadrature components carried are non-zero. The basic idea of our proposed generalization approach is to split the in-phase component and quadrature component into two independent components so that index modulation is applied independently on these two components, i.e., a subcarrier is not necessary to be active or inactive simultaneously for the in-phase and quadrature components.

For QPSK, the in-phase and quadrature component can be regarded as two independent QAM streams. If two independent index modulations are applied to these two independent QAM streams, the total number of combinations that can brepresented is given by $2^k C_n^K 2^k C_n^K G = 4^k C_n^k C_n^k G$ (32)

On the other hand, the total combination that OFDM-IM using QPSK represents is $4^k C_n^k C_n^k G$. As a result, we have

$$4^k C_n^k C_n^k G > 4^k C_n^k G \qquad (33)$$

For example, when n = 16;K= 10, the total number of bits that our scheme can transmit is given by

$$\left(\left(\lfloor \log_2 2^{10} C_{16}^{10} \rfloor\right) + \left(\lfloor \log_2 2^{10} C_{16}^{10} \rfloor\right)\right) G = 44G \qquad (34)$$

While the total numbers of bits that OFDM-IM and OFDM can transmit are both

$$\left(\lfloor \log_2 4^{10} C_{16}^{10} \rfloor\right) G = (\lfloor \log_2 4^n \rfloor) G = 32G \qquad (35)$$

In other words, when n = 16;K= 10, our proposed OFDM-GIM2 offers a 37.5% higher spectral efficiency compared to OFDM-IM. To successfully implement our proposed scheme, several changes are necessary and will be introduced in the following two subsections.

A) Revised Index Modulation Block:

Different from OFDM-IM, at the transmitter, the input bit strings allocated to each subblock are equally split into two parts, one for in-phase components' index modulation and the other for quadrature components' index modulation. The outputs of these two index modulation are then combined and constructed into a complex M-ary constellation symbol, as shown in Fig. 3. The output of these two index modulations are allocated the same total power such that the combined complex M-ary constellation symbol vector x still satisfies E{xH x}= N. At the receiver, the revised index demodulation block will simply do the inverse procedures of the revised index modulation block.

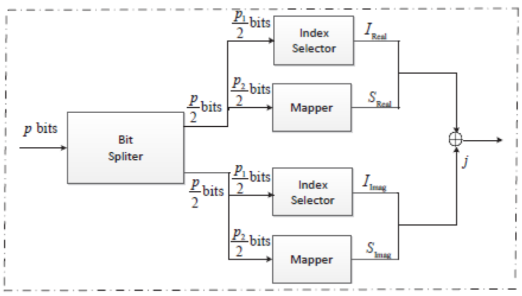


Fig 5: The generalized index modulation block of a subblock for OFDMGIM2 using QAM

Revised LLR detector:

The revised LLR detector for OFDM-GIM2 should have zero-forcing equalization first. For the α-th frequency domain received signal Y (α), for α= 1; : : : ;N, let

$$\mathbf{Y}^{"}(\boldsymbol{\alpha}) = \frac{\mathbf{Y}(\boldsymbol{\alpha})}{\mathbf{H}(\boldsymbol{\alpha})} \qquad (36)$$

Note that the zero-forcing equalization amplifies the noise power so the noise power in (11) should be changed accordingly by introducing a factor of H2(_). After equalization, what the detectors will detect aresm2 S instead of Hsm. Therefore, the revised LLR detector for the in-phase components is given by

$$\boldsymbol{\lambda}_{\mathbf{I}}(\boldsymbol{\alpha}) = \ln(\mathbf{K}) - \ln(\mathbf{n}-\mathbf{K}) + \frac{\mathbf{H}^2(\boldsymbol{\alpha})|\mathrm{Re}(\mathbf{Y}'(\boldsymbol{\alpha}))|^2}{\mathbf{N}_{\mathbf{0,F}}} + \ln(\sum_{\mathbf{m=1}}^{\mathbf{M}} \exp(-\frac{\mathbf{H}^2(\boldsymbol{\alpha})}{\mathbf{N}_{\mathbf{0,F}}}|\mathbf{Re}\left(\mathbf{Y}'(\boldsymbol{\alpha})\right) - \mathbf{s}_{\mathbf{m}}|^2)) \qquad (37)$$

where Re(Y’(α)) returns the real part of Y’(α). And the revised LLR detector for the quadrature components is given by

$$\boldsymbol{\lambda}_{\mathbf{Q}}(\boldsymbol{\alpha}) = \ln(\mathbf{K}) - \ln(\mathbf{n}-\mathbf{K}) + \frac{\mathbf{H}^2(\boldsymbol{\alpha})|\mathbf{Imag}(\mathbf{Y}'(\boldsymbol{\alpha}))|^2}{\mathbf{N}_{\mathbf{0,F}}} + \ln(\sum_{\mathbf{m=1}}^{\mathbf{M}} \exp(-\frac{\mathbf{H}^2(\boldsymbol{\alpha})}{\mathbf{N}_{\mathbf{),F}}}|\mathbf{Imag}(\mathbf{Y}'(\boldsymbol{\alpha}\mathbf{0}) - \mathbf{s}_{\mathbf{m}}|^2))$$

38)

Where Imag(Y’(α). returns the imaginary part of Y’(α). After that, the two sets of LLRs are independently fed to the inverse index modulation block to get an estimate of the input bit string.

IV. OFDM WITH ENHANCED SUBCARRIER USING QAM

In this paper, a novel orthogonal frequency division multiplexing (OFDM) scheme, called OFDM with index modulation (OFDM-IM), is proposed for operation over frequency-selective and rapidly time-varying fading channels. In this scheme, the information is conveyed not only by M-ary signal constellations as in classical OFDM, but also by the indices of the subcarriers, which are activated according to the incoming bit stream. Different low complexity transceiver structures based on maximum likelihood detection or log-likelihood ratio calculation are proposed and a theoretical error performance analysis is provided for the new scheme operating under ideal channel conditions. Then, the proposed scheme is adapted to realistic channel conditions such as imperfect channel state information and very high mobility cases by modifying the receiver structure. The approximate pair wise error probability of OFDM-IM3 using QAM is derived under channel estimation errors. For the mobility case, several interference unaware/aware detection methods are proposed for the new scheme. It is shown via computer simulations that the proposed scheme achieves significantly better error performance than classical OFDM due to the information bits carried by the indices of OFDM subcarriers under both ideal and realistic channel conditions.At the receiver ,the index demodulation block will simply do the inverse procedures of the generalized index modulation block.

V. IMPLEMENTATION COMPLEXITY ANALYSIS

Using a total power constraint is valuable for network planning as it allows to control the total power consumed in the whole network. Moreover, such a total transmit power constraint can provide a guideline for setting individual relay powers. when applying SNR balancing to the case of time-synchronous or time-asynchronous two-way relay networks is reasonable to assume that each relay, on average4, consumes fraction of the half of the total available power. This argument is particularly correct when the relays are moving randomly in the environment. In such a scenario, different relay channels appear to be drawn from the same i.i.d. distribution. Another scenario where a total power consumption is useful, is the case where the two transceivers and the relay nodes are powered up through the electric grid (implying that they are stationary). In this case, one has to ensure that the total power consumed by the network is restricted. Note also designing two-way relaying schemes under per-node power constraint is rather challenging. Indeed, even in the case of synchronous single-carrier two-way relay networks, optimal AF schemes can be computationally prohibitive as it amounts to a two-dimensional search on a sufficiently fine grid in the two-dimensional space of transceiver powers and solving a convex feasibility programming on each vertex of this grid. A synchronism and multi-carrier nature of the scheme considered in this paper only add to the level of difficulty associated with using per-node power constraints in the context of two-way relay networks. Moreover, in two-way relay networks, for any given channel realization, the solution to the sum-rate maximization problem under per-node power constraints may not be power-efficient in the sense that some of the relays may not consume all the power available .As a result, the average power consumed by the relays for different channel realizations will be less than the power available to each of them. For all these reasons, total power constraints have been adopted in the literature for performance analysis and optimal design. It is also worth mentioning that sum-rate maximization under a total power constraint provides an upper bound for the case when this maximization is performed under per-node power constraint s5. The extension to individual power constraints is an interesting problem especially in relay networks where relays are expected to be devices with a limited amount of power. This problem could be an interesting direction for future research on asynchronous two-way relay networks

$$c(\mathbf{w}) \triangleq \frac{\sigma^2(P_T - N\sigma^2\mathbf{w}^H\mathbf{w})}{(\mathbf{w}^H\mathbf{D}_1\mathbf{w}+\sigma^2)(\mathbf{w}^H\mathbf{D}_2\mathbf{w}+\sigma^2)}.$$

is easy to show that at the optimum, the second constraint in must be satisfied with equality. Otherwise, the optimal value of can be scaled up, thereby increasing the objective function in without violating the other constraints, and thus, contradicting optimality. Summarizing this subsection, we presented the problem of finding the rate region as the optimization problem. In the next subsection, we consider a relaxed version of the optimization problem and simplify the relaxed problem and finally solve it in the subsequent subsection.

Relaxing the Optimization Problem:We will show that for any feasible value of in the optimization problem, the solution to satisfies the last two constraints in and hence, for any feasible value of in the optimization problem any solution to is a solution. we define the Lagrangian corresponding to the inner maximization

$$\gamma_{1i} = \frac{|\mathbf{a}_i^H\mathbf{w}|^2}{\lambda_2} - 1, \quad \text{for } i = 1, 2, \ldots, N$$
$$\gamma_{2i} = -\frac{\lambda_1}{\lambda_2}|\mathbf{a}_i^H\mathbf{w}|^2 - 1, \quad \text{for } i = 1, 2, \ldots, N.$$

Solution to the Original Optimization Problem:

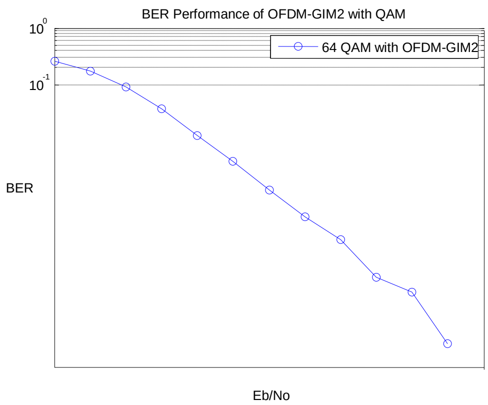
Now show that any solution to the optimization problem satisfies the constraints and , for any feasible value of , and thus, it is a solution to the optimization problem

$$= \frac{\sigma^2\left(P_T/N - \sigma^2\mathbf{w}_{n^o}^{o,H}\mathbf{w}_{n^o}^{o}\right)\left|\mathbf{b}_{n^o}^{H}\mathbf{w}_{n^o}^{o}\right|^2}{N\left(\mathbf{w}_{n^o}^{o,H}\mathbf{Q}_1^{(n^o)}\mathbf{w}_{n^o}^{o}+\sigma^2\right)\left(\mathbf{w}_{n^o}^{o,H}\mathbf{Q}_2^{(n^o)}\mathbf{w}_{n^o}^{o}+\sigma^2\right)}.$$

Holds true for any solution to the optimization problem regardless of the value. We show in the Appendix that for any feasible value of in the optimization problem, the optimal value of obtained by solving satisfies

VI. SIMULATION RESULT

The simulation results of the proposed schemes are shown and compared with that of OFDM-IM under frequency selective channels. In all simulations, we assumed the same system parameters as in [1] and [23], i.e., N = 128; V = 10 and L = 16: The SNR is defined as Eb=N0;T , where Eb = (N + L)=B is the average transmitted energy per bit. The BER performance of these schemes was evaluated via Monte Carlo simulations. Fig. 4 shows the BER performances of the OFDM-GIM1 schemes with two different K sets, classical OFDM and OFDM-IM for BPSK. The subblock size n is set to 8. When K = f3; 5g, according to (16), it is known that 11 bits can be transmitted per subblock. However, for OFDM-IM with n = 8, only 10 bits can be transmitted per subblock. Our proposed OFDM-GIM1 scheme achieves 10% higher spectral efficiency [1] at the cost of a BER performance loss lower than 0.5 dB. When K = f1; 2; 3; 4; 5; 6g, according to (16), 12 bits can be transmitted per subblock, and thus a 20% higher spectral efficiency is achieved at the cost of a BER performance loss up to 2.5 dB. The 2.5 dB BER performance loss was observed at the BER probability of 10^3.



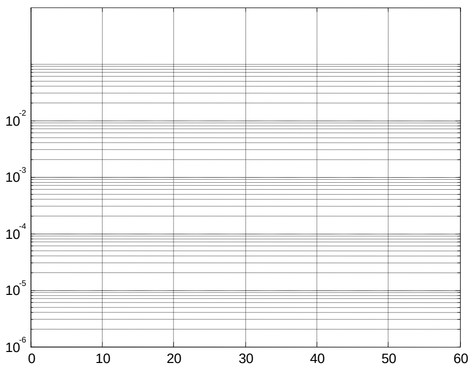


Fig.6. Ber performanace of OFDM-IIM and OFDM-IGIM3,QAM

VII. CONCLUSION:

In this paper, two generalization schemes of OFDM-IM are presented. To implement these two schemes, generalized index modulation blocks and upgraded LLR detectors are proposed, respectively. Interleaving is introduced to improve the BER performance of our proposed schemes in low SNR region. Both generalization schemes achieve higher spectral efficiency than OFDM-IM. When the same spectral efficiencies are considered, our proposed generalization schemes show consistent BER performance gain in all SNR regions. We demonstrated that the two generalization schemes are compatible with each other and their combined scheme greatly outperforms existing works in spectral efficiency and BER performance, at the cost of a little higher complexity.

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