A Unified Structure for Multi-Carrier Modulations in Power-Line Communications

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Abstract—The power line channel is characterized by a multi-path behavior. A direct consequence is that Multi-Carrier Modulation (MCM) is generally retained for Power Line Communication (PLC). In addition to OFDM, or equivalently DMT, Wavelet OFDM and Hermitian Symmetry OFDM/OQAM (HS-OQAM) have been recently proposed for PLC transmission. In this paper we propose a unified and efficient modulation scheme that can generate all these different modulation schemes.

I. INTRODUCTION

Power lines are now considered as an attractive medium to transmit information. Indeed, the electrical network is present in many places, so one can avoid the deployment of new infrastructures. However, one of the drawbacks is that transmission over power lines suffers from many impairments. Consequently, it makes the PLC channel a frequency selective behavior. In order to fight against frequency selectivity, Multi-Carrier Modulation (MCM) tends to be a good solution.

Orthogonal Frequency Division Multiplexing (OFDM) is now a widely used MCM technique to counteract multi-path fading channels. In PLC specifications, DMT is proposed as a physical modulation technique. It is essentially an OFDM-based modulation with the Hermitian Symmetry (HS) entries [1]. However, an important drawback of OFDM, a equivalent of DMT, is that its rectangular pulse shaping only provides a minimum stopband attenuation of 13 dB. Since then, a large number of studies have been done in order to find the possible advanced MCM schemes that could provide better frequency selectivity, or time-frequency localization. Consequently, one proposal addresses Wavelet OFDM (WOFDM) which is based on the Cosine Modulated Filter Bank (CMFB) and with entry symbols that are multi-level (Pulse Amplitude Modulation) PAM [2]. Hence, same as DMT, the base-band transmit signal is real-valued. Besides WOFDM, Siohan et al. [3] proposed an OFDM offset QAM, named here OQAM in short, in radio transmission context where the data symbols transmitted over each sub-carrier correspond to real PAM symbols that are issued from a complex QAM constellation. As the duration of these PAM symbols are half the one of the QAM ones, we get the same spectral efficiency as for OFDM transmitting QAM symbols without using Cyclic Prefix (CP). Both WOFDM and OQAM modulations can have a pulse shaping providing a good localization in time and frequency. Moreover, the utilization of OQAM in PLC context becomes possible with its HS version [4] leading to an HS-OQAM system.

As to the implementation issue, Chan et al. proposed an efficient implementation algorithm for DMT based on Fast Cosine/Sine transforms (FCT/FSTs) [5]. The idea, inspired by Lee’s cosine transform algorithm [6], takes advantage of the symmetry property of DMT entries to reduce the complexity up to around 80% of the conventional DMT implementation with Fast Fourier Transforms (FFTs).

In this paper, we aim at extending Chan’s idea to HS-OQAM and WOFDM implementation with FCT/FSTs. Ultimately, we come up with a unified multi-carrier modulator whereby, we can arbitrarily switch the above three modulation techniques from one to the other. The reasons of proposing this unified structure are: In the first, based on our previous work [7], wherein we gave an analytical discussion on HS-OQAM vs. DMT in terms of their capacities and we unveiled a selection threshold for judging when we shall choose HS-OQAM or DMT. Therefore, our proposed structure can tell us how to practically realize this modulation selection. In the second, as reported in IEEE P1901 (draft standard for broadband over power line networks) [8], the proposals to the dual physical layer options have been the subject of much controversy, i.e. one option: DMT modulation and the other option: WOFDM modulation. Thus, our unified modulator can include all of these existing modulations plus HS-OQAM. In Sec. II, we briefly remind of Chan’s implementation for DMT modulation. In Sec. III, we present the efficient implementation for HS-OQAM modulation with FCT/FSTs. In Sec. 5, we demonstrate the WOFDM implementation with FCT/FSTs. In Sec. V, we show our final unified modulator structure.

For the notations, we first define four types of fast cosine transform kernel matrices, for $k, n = 0, \ldots, N - 1$, as [6]

$$\begin{align*}
[FCT_I]_{k,n} &= C_{2N}^{kn} = \cos \left( \frac{kn}{2N} \right), \\
[FCT_{II}]_{k,n} &= C_{2N}^{k(n+\frac{1}{2})} = \cos \left( \frac{k(n+\frac{1}{2})}{2N} \right), \\
[FCT_{III}]_{k,n} &= C_{2N}^{(k+\frac{1}{2})n} = \cos \left( \frac{(k+\frac{1}{2})n}{2N} \right), \\
[FCT_{IV}]_{k,n} &= C_{2N}^{(k+\frac{1}{2})(n+\frac{1}{2})} \nonumber \\
&= \cos \left( \frac{(k+\frac{1}{2})(n+\frac{1}{2})}{2N} \right),
\end{align*}$$

where $k$ stands for the entry index and $n$ denotes the output index. If we replace the cosine operation with sine, it results
in a set of sine transform kernel matrices.

II. DMT MODULATION

In this section, we give a brief remind of Chan’s implementation for DMT modulation. Assume that we have 2M sub-carriers, the Hermitian Symmetry (HS) constraints for DMT yield \( c_{0,n} = c_{M,n} = 0 \) and \( c_{m,n} = c_{2M-m-n} \) for \( m = 1, \ldots, M - 1 \), where \( c_{m,n} \) are complex-valued symbols mapped to \( m \)-th sub-carrier on the time index \( n \). The modulated DMT signal can be expressed as

\[
u[k] = \frac{1}{2M} \sum_{n \in \mathbb{Z}} \sum_{m=0}^{2M-1} c_{m,n} W^{-mk}, \tag{2}\]

where \( W \) is the Fourier transform kernel denoting \( W = e^{-j \frac{2\pi}{2M}} \). Taking into account the HS constraints and decomposing (2) leads to

\[
u[k] = \frac{1}{M} \sum_{n} \left[ \sum_{m=0}^{M-1} x_{m,n} c_{2M}^{mk} - \sum_{m=0}^{M-1} x_{m,n} c_{2M}^{-mk} \right]
= \frac{1}{M} \left[ \text{FCTI}(k) - \text{FSTI}(k) \right], \tag{3}\]

where \( x_{m,n}^r \) and \( x_{m,n}^i \) are the real and imaginary part of \( c_{m,n} \), respectively. Thus, the DMT modulation (3) can be implemented based on FCT/FST type I (FCT/FST-I) kernel. Further, the symmetry properties of FCT/FST-I yield

\[
\begin{align*}
\text{FCTI}(k) &= \text{FCTI}(2M - k) \quad \text{for} \quad k = 0, \cdot \cdot \cdot, M - 1 \\
\text{FSTI}(k) &= -\text{FSTI}(2M - k) \quad \text{for} \quad k = 0, \cdot \cdot \cdot, M - 1 \\
\text{FCTI}(M) &= \sum_{m=0}^{M-1} x_{m,n}^r (-1)^m \quad \text{for} \quad M = 0, 2M - 1 \\
\text{FSTI}(M) &= 0
\end{align*}\tag{4-7}\]

These properties allow us to calculate only \( M \)-point FCT/FST-I for obtaining in total \( 2M \)-point results. Next, we briefly demonstrate the efficient implementation of FCT/FST-I given in [5]. For the \( M \)-point FCT-I, we have

\[
\text{FCTI}(l) = \sum_{m=0}^{M-1} x[m] c_{2M}^{ml}, \quad \text{for} \quad l = 0, \ldots, M - 1. \tag{8}\]

Next, based on Lee’s algorithm [6], (9) can be expressed as

\[
\begin{align*}
\text{FCTI}(l) &= \sum_{m=0}^{M/2-1} x[2m] c_{2M}^{ml} + \sum_{m=0}^{M/2-1} x[2m+1] c_{2M}^{(2m+1)l} \\
&\quad + \frac{1}{2C_{2M}^l} \left( \sum_{m=0}^{M/2-1} x[2m+1] + x[2m-1] \right) C_{2M}^m h(l) \\
&\quad + \frac{x[M-1]}{q(l)} (-1)^l, \quad \text{for} \quad l = 0, \ldots, M - 1. \tag{9}\end{align*}\]

where we assume \( x[-1] = 0 \). For the rest half FCT-I outputs we have

\[
\text{FCTI}(M - l) = g(l) - \frac{1}{2C_{2M}^l} [h(l) + q(l)]. \tag{10}\]

The special case FCTI(\( M/2 \)) yields

\[
\text{FCTI}\left( \frac{M}{2} \right) = \sum_{m=0}^{M-1} x[m] \cos \frac{m\pi}{2}. \tag{12}\]

As (10) shows that the \( M \)-point FCT-I can be decomposed into two \( M/2 \)-point FCT-Is. Then we can apply the technique of divide-and-conquer to recursively expand the \( M \)-point FCT-I until 1-point FCT-I. The similar algorithm is found for FST-I transform implementation. The details can be found in [5]. Finally, the implementation structure of DMT is depicted in Fig. 1, where the block PRP stands for PRe-Processing stage which is in charge of taking the real and imaginary parts of the entry symbols and separately mapping them to the FCT/FST-I inputs; POP block denotes PPost-Processing stage which mainly deals with the FCT/FST-I outputs w.r.t (3) and (4)-(7). Furthermore, the CP appending and windowing processes [1] are applied in this block; The P/S block gives a parallel to serial operation.

III. HS-OQAM MODULATION

A FCT/FST-I based DMT implementation was introduced in the preceding section. In this section, we extend to HS-OQAM modulation. The discrete-time modulated HS-OQAM signal yields [4]

\[
u[k] = \sum_{m=0}^{2M-1} \sum_{n \in \mathbb{Z}} a_{m,n} \exp \left[ j \left( \frac{2\pi}{2M} (m-\frac{1}{2}) + \phi_{m,n} \right) \right], \tag{13}\]

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with HS constraints given, for \( m = 1, \ldots, M - 1 \), by

\[
a_{0,n} = a_{M,n} = 0; \\
a_{m,n} = a_{2M-m,n}(-1)^{D-M-n}e^{-j2\phi_0},
\]

(14)

where \( a_{m,n} \) are real-valued symbols obtained by taking the real or imaginary parts of the complex-valued symbols; \( p[k] \) is the prototype filter; \( L \) is its length and we set \( D = L-1 \); \( \phi_{m,n} \) is an additional phase term and can be generally expressed as \( \phi_{m,n} = \frac{\pi}{2}(m+n) + \phi_0 \), with \( \phi_0 \) being arbitrarily chosen, e.g. \( \phi_0 = -\pi mn \) in [3]. After a simple mathematical combination, (13) can be alternatively expressed as

\[
u[k] = \sum_{m=0}^{2M-1} \sum_{n \in \mathbb{Z}} x_0^m[n]f_m[k-nM],
\]

(15)

where \( x_0^m[n] = a_{m,n}j^n \) and its phase rotation term \( j^n \) permits to have the staggered offset QAM structure [3]. The synthesis filters form as \( f_m[k] = p[k]e^{j\frac{\pi}{2}m(k-D/2)} \). The HS-OQAM transmitter is depicted in Fig. 2.

![HS-OQAM transmitter structure](image)

Fig. 2. HS-OQAM transmitter structure.

Taking the Z-transform of \( u[k] \) yields

\[
U(z) = \sum_{m=0}^{2M-1} X_0^m(z^M)F_m(z),
\]

(16) results in

\[
U(z) = \sum_{l=0}^{2M-1} z^{-(2M-1-l)} \times \left[ G_l(z^2) \sum_{m=0}^{2M-1} X_0^m(z)W^{m(D-M)}W^{-ml'} \right]_{T_l(z)} \]

(19)

where \( l' = 2M - 1 - l \); \( [\cdot]_M \) denotes expander operation with factor \( M \). Next, taking advantage of (14) with \( \phi_0 = -\pi mn \), for \( m = 1, \ldots, M - 1 \), results in

\[
X_0^{2M-m}(z) = Z\{x_0^{2M-m}[n]\} = (-1)^{M-D}A_m(jz)
\]

(20)

Thus, we can further decompose \( T_l(z) \) in (19) as

\[
T_l(z) = \sum_{m=0}^{M-1} \left[ X_0^m(z)W^{m(D-M)}W^{-ml'} \right] + X_0^{2M-m}(z)W^{(2M-m)(D-M)}W^{-(2M-m)l'}
\]

\[
= 2 \sum_{m=0}^{M-1} \Re\left\{A_m(-jz)W^{m(D-M)}W^{-ml'}\right\}
\]

(21)

Next, we define \( X_m(z) = A_m(-jz)W^{m(D-M)} = X_0^m(z) + jX_0^m(z) \), where \( X_0^m \) and \( X_0^m \) are the real and imaginary part of \( X_m(z) \), respectively. Then, we find a FCT/FST-I based implementation for (21) as for \( l' = 0, \ldots, 2M - 1 \),

\[
T_l(z) = 2 \sum_{m=0}^{M-1} \left[ X_0^m(z)C_{M-1}^{ml'} - X_0^m(z)S_{M-1}^{ml'} \right]
\]

\[
= 2[\text{FCT}_1(l') - \text{FST}_1(l')].
\]

(22)

Substituting (22) into (19), we have the expression of FCT/FST-I based HS-OQAM modulation

\[
U(z) = 2 \sum_{l=0}^{2M-1} z^{-(2M-1-l)} \left[ G_l(z^2) \times [\text{FCT}_1(l') - \text{FST}_1(l')]\right]_M
\]

(23)

![Efficient FCT/FST-I based blocks for HS-OQAM modulation](image)

Fig. 3. Efficient FCT/FST-I based blocks for HS-OQAM modulation.

Similar to DMT modulation, in Fig. 3, we show the implementation structure of HS-OQAM modulation using FCT/FST-I. The PRP block calculates \( X_m(z) \) then separates the real and imaginary parts, \( X_0^m(z) \) and \( X_0^m(z) \), respectively; The POP block computes the term \( T_l(z) \) w.r.t (22) and again (4)-(7) using the FCT/FST-I outputs; The PF block denotes the
Polyphase Filtering process in (23). The PF block is nothing else than a diagonal matrix, $G_{\text{diag}}(z)$, with

$$G_{\text{diag}}(z) = \text{diag}[G_0(z), \ldots, G_{2M-1}(z)].$$

Note that, if the prototype filter has the length equal to symbol duration $2M$, then for each carrier, the polyphase filter has only one coefficient. Further, different to DMT, the P/S is parallel to serial block with up-sampling factor of $M$ depicted in Fig. 4. However, for DMT modulation, this up-sampling factor is $2M$ rather than $M$.

![Fig. 4. HS-OQAM P/S block.](image)

IV. WOFDM MODULATION

Besides HS-OQAM and DMT modulations, Koga et al. proposed a WOFDM modulation for PLC [2]. The general transmitter structure is depicted in Fig. 5. We find that the number of sub-carriers of WOFDM is half of the DMT and HS-OQAM ones. Since entry symbols of WOFDM are multi-level PAM and the synthesis filters of WOFDM are cosine modulated kernel, therefore, WOFDM can directly generate a real-valued baseband modulated signal without any HS constraints. In this section, we derive the implementation algorithm for WOFDM with FCT/FSTs.

![Fig. 5. WOFDM transmitter structure.](image)

The discrete-time modulated WOFDM signal yields [10]

$$u[k] = \sum_{m=0}^{M-1} \sum_{n \in \mathbb{Z}} a_{m,n} f_m[k-nM],$$

with the synthesis filters

$$f_m(k) = 2p_0(k) \cos \left( \frac{\pi}{M} \left( m + 0.5 \right) \left( k - \frac{L-1}{2} \right) - \theta_m \right),$$

where $a_{m,n}$ are multi-level PAM symbols; $\theta_m = (-1)^m \frac{\pi}{2}$; $p_0(k)$ is the prototype filter and $L$ is the length of $p_0(k)$, we can also simply set $D = L-1$. Note that WOFDM can use the same prototype filter as HS-OQAM one. Because they have the same perfect reconstruction conditions [4].

The modulated WOFDM signal (24) can be re-written as

$$u[k] = 2R \left\{ \sum_{m=0}^{M-1} \sum_{n \in \mathbb{Z}} a_{m,n} f'_m[k-nM] \right\},$$  

where $f'_m(k) = e^{j\left( \frac{\pi}{M} (m+0.5)(k-\frac{L}{2}) - \theta_m \right)}$. Then, taking the Z-transform of (25) leads to

$$U(z) = 2R \left\{ \sum_{m=0}^{M-1} A_m(z^M) F'_m(z) \right\},$$

with

$$F'_m(z) = W^{(m+0.5)\frac{L}{2}} e^{-j\theta_m} P_0 \left( zW^m + \frac{1}{2} \right).$$

(27) can be compactly written as (polyphase type 2) [9]

$$F'_m(z) = \sum_{l=0}^{2M-1} z^{-(2M-1-l)} R_{l,m}(-z^{2M}),$$

with

$$R_{l,m}(-z^{2M}) = W^{(m+0.5)(\frac{L}{2} - l')} e^{-j\theta_m} G_{l'}(-z^{2M}),$$

where $l' = 2M - 1 - l$. Substituting (28) into (26) results in

$$U(z) = 2R \left\{ \sum_{l=0}^{2M-1} z^{-(2M-1-l)} \sum_{m=0}^{M-1} X_m(z) W^{-(m+\frac{1}{2})l'} G_{l'}(-z^{2M}) \right\},$$

where $X_m(z) = W^{(m+0.5)\frac{L}{2}} e^{-j\theta_m} A_m(z) = X_m^*(z) + jX_m^t(z)$. Let us further decompose $W^{-(m+\frac{1}{2})l'} = C_2^{(m+\frac{1}{2})l'} + j X_2^{(m+\frac{1}{2})l'}$, (29) can be re-written, for $l' = 0, \ldots, 2M - 1$, as

$$U(z) = 2 \sum_{l=0}^{2M-1} z^{-(2M-1-l)} \left[ G_{l'}(-z^{2M}) \sum_{m=0}^{M-1} X_m(z) C_2^{(m+\frac{1}{2})l'} - \sum_{m=0}^{M-1} X_m(z) X_2^{(m+\frac{1}{2})l'} \right],$$

$$= \sum_{l=0}^{2M-1} z^{-l'} \left[ G_{l'}(-z^{2M}) \left( 2 (\text{FCT}^{(l')} - \text{FST}^{(l')}) \right) \right],$$

where $\text{FCT}^{(l')} = 2 \sum_{m=0}^{M-1} X_m(z) C_2^{(m+\frac{1}{2})l'}$ and $\text{FST}^{(l')} = 2 \sum_{m=0}^{M-1} X_m(z) X_2^{(m+\frac{1}{2})l'}$.
In the HS-OQAM case, we have a PRP block. The PRP block uses FCT/FST type III (FCT/FST-III) kernel. The implementation block is shown in Fig. 7, where the PRP block is the same as that of HS-OQAM case. The P/S block is the same as that of HS-OQAM modulation.

Thereby, from (30), we find that the WOFDM modulation can use FCT/FST type III (FCT/FST-III) kernel. The properties of FCT/FST-III kernels yield
\[
\begin{align*}
\text{FCT}_{\text{III}}(k) &= -\text{FCT}_{\text{III}}(2M - k) \quad (31) \\
\text{FST}_{\text{III}}(k) &= \text{FST}_{\text{I}}(2M - k) \quad (32) \\
\text{FCT}_{\text{III}}(M) &= 0 \\
\text{FST}_{\text{III}}(M) &= \sum_{m=0}^{M-1} X_m^r(z)(-1)^m \quad (34) \\
\text{FCT}_{\text{III}}(0) &= \sum_{m=0}^{M-1} X_m^r(z) \quad (35) \\
\text{FST}_{\text{III}}(0) &= 0 \quad (36)
\end{align*}
\]

The implementation block is shown in Fig. 7, where the PRP block calculates the term \( X_m(z) = W^{m+0.5} e^{-j\theta_m} A_m(z) \), and then, separates its real and imaginary parts; The POP block computes the term \( T_l(z) \) w.r.t (30) and (31)-(36). Similar to HS-OQAM case, we have a PF block. The PF block has a diagonal matrix form as
\[
\text{G}_{\text{diag}}(-z) = \text{diag}[G_0(-z), \ldots, G_{2M-1}(-z)],
\]
and for each sub-carrier \( i \), for \( i = 0, \ldots, 2M - 1 \), \( G_i(-z) \) has only one coefficient when the prototype length is equal to \( 2M \). Further, in this case, we find the fact that \( \text{G}_{\text{diag}}(-z) \) is identical to the \( \text{G}_{\text{diag}}(z) \) which is the polyphase matrix of HS-OQAM case. The P/S block is the same as that of HS-OQAM modulation.

In order to have a unified transform kernel, we now show the link between the FCT/FST-III and -I, such that the WOFDM modulation can be implemented with FCT/FST-I. The formulation of FCT-III can be expressed as
\[
y(l) = \sum_{m=0}^{M-1} x[m] C_{2M}^{(m+0.5)l} = \sum_{m=0}^{M-1} x[m] \tilde{C}_{2M}^{(2m+1)l}, \quad (37)
\]
where we denote \( \tilde{C}_{2M}^{(2m+1)l} = \cos \left( \frac{\pi}{2M} ml \right) \). By applying the trigonometric identity as
\[
2 \tilde{C}_{2M}^{(2m+1)l} \tilde{C}_{2M}^{(2M+1)l} = \tilde{C}_{2M}^{(2m)l} + \tilde{C}_{2M}^{(2m+2)l}, \quad (38)
\]
then (37) can be written as
\[
y(l) = \frac{1}{2 \tilde{C}_{2M}^{(2m+1)l}} \left( \sum_{m=0}^{M-1} x[m] \tilde{C}_{2M}^{(2m)l} + \sum_{m=0}^{M-1} x[m] \tilde{C}_{2M}^{(2m+2)l} \right)
\]
\[
= \frac{1}{2 \tilde{C}_{2M}^{(2m+1)l}} \left( x[M-1](-1)^l + \sum_{m=0}^{M-1} (x[m] + x[m-1]) \tilde{C}_{2M}^{(2m+1)l} \right). \quad (39)
\]
Since \( \tilde{C}_{2M}^{(2m)l} = C_{2M}^{(2ml)} \), which is actually the transform kernel of \( M \)-point FCT-I. Thereby, \( M \)-point FCT-III can be realized based on FCT-I. The FCT-III can be implemented by FCT-I as shown in Fig. 6(a). A similar process can be applied to
Then (40) can be written as

\[ y(l) = \frac{1}{2C_2^M} \left( \sum_{m=0}^{M-1} x[m] \tilde{S}_2^{(2m+1)l} + \sum_{m=0}^{M-1} x[m] \tilde{M}_2^{(2m+1)l} \right) \]

(41)

Again, since \( \tilde{S}_2^{(2m)l} = \tilde{S}_2^{(2m+1)l} \), which is the transform kernel of \( M \)-point FST-I, the FST-III can be implemented by FST-I as shown in Fig. 6(b).

Furthermore, in Fig. 6(a),6(b), the process before the entry of FCT/FST-I and the one after the output of FCT/FST-I can be directly put into the PRP and POP blocks, respectively.

V. UNIFIED STRUCTURE FOR MCMs

In the previous sections, we presented the FCT/FST-I based implementations for DMT, HS-OQAM and WOFDM, respectively. Since they share a same transform kernel, thus, a unified modulation can be expected. The unified modulator structure is shown in Fig. 8. This unified modulator consists of three parts: PRP part, transform part and POP part. In the PRP part, it includes three blocks. Each block is in charge of executing the pre-processing of a selected modulation type. In the POP part, it also has three blocks. Each block is responsible for the process after the FCT/FST-I transform stage, i.e. for DMT modulation type, the block corresponds to the POP plus P/S processes of DMT; For HS-OQAM or WOFDM modulation type, the block corresponds to POP plus PF plus P/S processes. Moreover, a MODulation SELection (MODSEL) parameter is introduced in Fig. 8 to choose the desired modulation type among DMT, HS-OQAM and WOFDM. The decision rule of selecting the proper scheme can be referred to, for instance, the maximum throughput criterion which has been detailed in [7]. Moreover, it is worth noting that this unified MCM modulator can be applied not only to generate above three modulations but also the other MCM modulation techniques such as Filter Multi-Tone (FMT).

VI. CONCLUSION

In this paper, we extend a low complexity implementation algorithm to HS-OQAM and WOFDM for PLC using FCT/FSTs. Furthermore, a unified modulator is proposed. This can be seen as a block box modulator by which we can choose to generate a large set of the existing MCM modulation techniques of PLC.

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