A measurement method of the DTMB modulator

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Abstract—In this paper, a measurement method for DTMB modulator is proposed, which performs the estimation of timing and frequency offsets, channel impulse response (CIR), as well as modulation error ratio (MER) by using only two consecutive frames in an iterative manner. Both simulation and lab testing results show that the measurement method could reflect the performance of DTMB modulator with high accuracy.

Index Terms—DTMB modulator, measurement, two frames, iteration

I. INTRODUCTION

DTMB (Digital Television/Terrestrial Multimedia Broadcasting), adopted as the mandatory Chinese Digital Terrestrial Television Broadcasting (DTTB) standard since August 1, 2007 [1]–[3], has been attracted increasing attention. How to evaluate the quality of DTMB modulator is of the great importance for both industrial and academia.

The conventional receiver is designed to demodulate the transmitted data in real-time, using robust algorithms to combat the non-ideal effects introduced during the transmission, such as frequency selectivity caused by multipath delay spread and carrier frequency offset, etc., However, in the evaluation system, it is only necessary to overcome the effects of weak frequency selectivity, moderate timing and frequency offsets, and to evaluate the performance of the modulator precisely, since the receiver is connected to the modulator directly by a cable instead of actual wireless channels. Finally, the receiver dose not need to process data in real-time in the evaluation system, and the analytical module/unit can even be separated from the data collector.

At the physical (PHY) layer, measurements are usually performed on parameters like modulation error ratio (MER) [4]. In [5], [6], analysis and experiments are offered for the study of the performance of the DVB system, and measurements of MER are performed to verify the method. In [7], technical specifications and methods of measurements for DTMB modulator are provided where MER is also the most important parameter at PHY layer. MER might be measured with lower reading than the actual value without precisely compensating time/frequency offsets as well as channel frequency selectivity. Besides, these parameters themselves reflect the performance of different components of the modulator, and can help users for the modulator troubleshooting. The proposed method is able to measure the parameters including MER, time/frequency offsets and channel impulse response (CIR).

To ensure the measurement accuracy, a large number of consecutive frames are usually used to perform the parameter estimation, such that the conventional modulator evaluation system can only reflect the average performance over a long time. However, our proposed method uses only two consecutive frames in an iterative manner, which reflects the time-varying property of the modulator. The precision is guaranteed by filling up the transmitted data with same sequences, adopting joint iterative synchronization, as well as the iterative and data-aided channel estimation algorithms. In addition, the method with fewer frames enables remote measurement and rapid measurement.

The rest of the paper is organized as follows. Section II introduces the system architecture for the DTMB modulator. Section III describes two typical modulation modes supported by DTMB modulator and the corresponding measurement methods. In Section IV, the simulation and lab testing results are presented to evaluate the performance of DTMB modulator. In Section V, conclusions are drawn.

II. SYSTEM MODEL

DTMB is a block transmission system supporting both single and multi-carrier modulation with modulator schematic diagram shown in Fig. 1. The information source data is first passed to a forward error control (FEC) encoder after scrambling, the output binary sequence of which is mapped to M-ary quadrature amplitude modulation (M-QAM) symbols by the constellation mapper. The frame body processing, depending on single or multi-carrier mode, is described in detail in references [2], [3]. After the frame body (FB, noted as \(d[n]\)) processing, the symbols of length \(N_d\) are multiplexed with the frame header (FH, noted as \(c[p]\)) of length \(N_c\) to form a signal frame of length \(N = N_c + N_d\), as shown in Fig. 2. The signal frame is pulse-shaped by a square root raised cosine (SRRC) filter. Subsequently, the output signal of the SRRC filter is transmitted after the RF module.

In order to evaluate the performance of a DTMB modulator precisely, several non-ideal characteristics should be considered.

The first factor is WFS which may degrade the evaluation precision [8]. Instead of the severe frequency selectivity in real wireless channel, WFS only exists in the evaluation system, which is originated from the imperfect digital-to-analog (D/A) conversion and non-ideal filtering in the DTMB modulator.

The second factor is timing and frequency offsets. Since the measurement receiver has been designed carefully and can be considered as the perfect receiver, all the offsets are assumed being introduced by the modulator. The uncompensated offsets will dramatically degrade the precision of the evaluation system, thus the offsets should be estimated accurately and
compensated precisely. At the same time, the offsets are also important measurement parameters which reflect the frequency offsets of the sampling clock and the local oscillator in the DTMB modulator.

The last one is the dynamic performance in a short time. The components of the modulator usually have time-varying properties. For example, the temporal variations of the phase noises from local oscillator would degrade the performance [9]–[11] of the modulator and we must reflect the dynamic changes of the modulator precisely.

To solve the problems mentioned above, an iterative measurement method is proposed, which will be described in detail in the following section.

III. Measurement Method

The proposed iterative measurement method is carried out according to the following steps:

1) Signal capturing: The signal output from the DTMB modulator is transmitted through the cable and then passed through the frequency down-converter and the analog filter, to become low intermediate frequency signal. This signal is then sent to the analog-to-digital (A/D) convertor to obtain the sampled signals, which are used for performance evaluation. These procedures are all performed by the data collector, which could be physically separated from the evaluation system. The advantage is that if the capturing environment is severe, the data could be collected and then transmitted back to analyze.

2) Joint iterative synchronization: In order to reflect the performance of the DTMB modulator in a short time, we use only two frames to perform synchronization instead of several hundreds frames used in conventional systems. To obtain accurate estimations of all synchronization parameters with such limited data, an iterative method is introduced to guarantee the synchronization accuracy.

3) Iterative channel estimation and equalization: For the same reason mentioned above, an iterative channel estimation method is also proposed to compensate the effect of WFS.

4) Results output: Three groups of parameters are shown to reflect the performance of the DTMB modulator. The first group includes the conventional parameters used for evaluating performance of modulator, e.g., MER. The second one includes the timing and frequency offsets. The CIR and channel frequency response (CFR) obtained in step 3) are given as the last group of parameters.

The FH of the DTMB signal can be configured as 420, 595, or 945 symbols, known as PN420, PN595 and PN945, respectively. Since the FH generation methods of PN420 and PN945 are the same, in this section, the algorithms of iterative joint synchronization, channel estimation and equalization for PN420 and PN595 mode are presented in detail respectively.

A. Measurement method for PN420 mode

The frame structure of the DTMB signal using PN420 mode is shown in Fig. 3.

1) Joint iterative synchronization: As shown in Fig. 4, the iterative method is carried out in the following steps, which is used to perform synchronization precisely with only two
signal frames. After signal capturing, frame synchronization is performed to obtain frame start position. In addition, we could obtain the coarse estimation of sampling frequency offset (SFO), sampling phase offset (SPO) and carrier frequency offset (CFO). In the first iteration, time/frequency offsets estimated in step “frame synchronization” are compensated, and in the following, the offsets estimated in the previous iteration are compensated. After the compensation process, fine estimations of SFO, SPO and CFO are performed. As the transmission environment is much better than actual wireless channel, the iteration will converge within 3 to 4 iterations, and then we will obtain the compensated data and estimations of SFO, SPO and CFO as measurement parameters. The algorithm will be described as follows.

a) Frame synchronization

As pointed out in [12], the conventional frame synchronization algorithm [13] for DTMB system is very sensitive to CFO. An improved differential correlation method [14] is introduced in the evaluation system. The sampled received sequence \( y[n] \) can be represented as

\[
y[n] = \{c[n] * h[n]\}e^{j\omega n} + v[n]
\]  

(1)

where operator * denotes linear convolution, while \( h[n] \), \( v[n] \) and \( \omega \) denote CIR, additive white Gaussian noise (AWGN) and CFO, respectively.

The differential correlation is defined as

\[
R_{diff}[k] = \sum_{n=0}^{N-1} (y[n + k]y^*[n + k + D])(c[n]c^*[n + D])
\]

(2)

where the superscript * denotes the complex conjugation operator and \( D \) denotes the differential correlation operation. Although the initial phase of PN sequence varies in the PN420 mode, the distance between any two adjacent correlation peaks is unique and could be used to detect the initial phase of the PN sequence and the frame start position [15].

b) Estimations of SFO and SPO

In [16] the proposed approach is based on the observation that the impact of SFO is equivalent to a relative time shift between two consecutive cyclic blocks, and in [13] the impact of SPO is equivalent to a relative time shift between the received PN sequence and the local PN sequence. We adopt these approaches with modifications in this paper.

In [16], the transmitted signal \( x(t) \) is shown in Fig. 5, where the gray areas represent the cyclic block and its replication with the time interval of \( T_F \); \( t_0 \) and \( T_C \) are the start time and the duration of the cyclic block. For \( t \in (t_0, t_0 + T_C) \), \( x(t) = x(t + T_F) \), which denotes the cyclic property.

In empty packets mode of the modulator, every frame is filled up with the same data, and we could regard the \( j \)th frame as the cyclic block of the \( (j+1) \)th frame.

The auto-correlation is given by

\[
R_{auto}[k] = \sum_{n=0}^{N_d} \{y^*_d[n]\}y^*_d[n + k], \quad k \in [N - 4, N + 4]
\]

(3)

where \( y^*_d[n] \) denotes the received data block after the compensation in the \( i \)th iteration.

The asymmetry of \( R_{auto}[k] \) can be measured as

\[
d_{auto}[k] = |R_{auto}[k + 1] - R_{auto}[k - 1]| + \frac{1}{4}(|R_{auto}[k + 2] - R_{auto}[k - 2]|)
\]

(4)

By normalizing \( d_{auto}[k_{auto}] \) the estimation of relative SPO between the two frames can be presented as

\[
\tilde{d}_{auto}[k_{auto}] = \frac{d_{auto}[k_{auto}] - d_{auto}[k_{auto} + 1]}{\sum_{k=1}^{N_{auto}} |d_{auto}[k_{auto} - 1] - d_{auto}[k_{auto}]|}
\]

(5)

\[
\tilde{d}_{auto}[k_{auto}] = \begin{cases} 
\frac{d_{auto}[k_{auto}] - d_{auto}[k_{auto} - 1]}{|R_{auto}[k_{auto} + 1]|}, & \text{if} \quad |R_{auto}[k_{auto} + 1]| \geq |R_{auto}[k_{auto} + 2]| \\
\frac{d_{auto}[k_{auto} - 1] - d_{auto}[k_{auto}]}{|R_{auto}[k_{auto}]|}, & \text{otherwise}
\end{cases}
\]

where \( k_{auto} \) represents the location of auto-correlation peak point. The fine estimation of SFO is obtained as

\[
\hat{f}_{fine} = \frac{\tilde{d}_{auto}[k_{auto}]}{N}
\]

(6)

In [13], the estimation of SPO is obtained from estimating the relative time shift between the received PN sequence and the local PN sequence through cross-correlation. We improve the measurement of the asymmetry of cross-correlation like (4).

The cross-correlation is given by

\[
R_{cross}[k] = \sum_{n=0}^{N_c-1} c[n]y^*[n + k]
\]

(7)

The asymmetry of \( R_{cross}[k] \) can be measured as

\[
d_{cross}[k] = |R_{cross}[k + 1]| - |R_{cross}[k - 1]| + \frac{1}{4}(|R_{cross}[k + 2] - |R_{cross}[k - 2]|)
\]

(8)

Normalizing \( d_{cross}[k_{cross}] \) yields the estimation of SPO

\[
\tilde{d}_{cross}[k_{cross}] = \begin{cases} 
\frac{d_{cross}[k_{cross} - 1] - d_{cross}[k_{cross}]}{|R_{cross}[k_{cross} - 1]|}, & \text{if} \quad |R_{cross}[k_{cross} - 1]| \geq |R_{cross}[k_{cross} + 1]| \\
\frac{d_{cross}[k_{cross}] - d_{cross}[k_{cross} + 1]}{|R_{cross}[k_{cross}]|}, & \text{otherwise}
\end{cases}
\]

(9)

where \( k_{cross} \) represents the location of cross-correlation peak point.

In coarse estimation of step “frame synchronization”, we extend the approach based on the relative time shift to the differential correlation for coarse estimation.

The asymmetry of \( R_{diff}[k] \) can be measured as

\[
d_{diff}[k] = |R_{diff}[k + 1]| - |R_{diff}[k - 1]| + \frac{1}{4}(|R_{diff}[k + 2] - |R_{diff}[k - 2]|)
\]

(10)
Normalizing $d_{diff}[k_{diff}]$ yields the estimation of SPO shown as

$$\hat{\mu}_{\text{coarse}} = \begin{cases} 
    \frac{d_{diff}[k_{diff}] - d_{diff}[k_{diff} - 1]}{d_{diff}[k_{diff} + 1] - d_{diff}[k_{diff}]} & \text{if } |R_{diff}[k_{diff} - 1]| \geq |R_{diff}[k_{diff} + 1]| \\
    d_{diff}[k_{diff}] - d_{diff}[k_{diff} + 1] & \text{otherwise}
\end{cases}$$

(11)

where $k_{diff}$ represents the location of differential correlation peak point.

The relative SPO is measured as

$$\hat{\Delta}_{\text{coarse}} = k_{diff}' - k_{diff} - N + \hat{\mu}_{\text{diff}} - \hat{\mu}_{\text{coarse}}$$

(12)

where $k_{diff}$ and $k_{diff}'$ denote two adjacent correlation peak points respectively.

SFO can be obtained as

$$\hat{f}_{\text{coarse}} = \frac{\hat{\Delta}_{\text{coarse}}}{N}$$

(13)

c) Estimation of CFO

The more accurate algorithm has smaller estimation range; therefore three kinds of methods are used to estimate CFO:

- The coarse CFO estimation can be given by differential correlation [14], i.e.

$$\hat{\omega}_{\text{coarse}} = \frac{\text{arg}(R_{diff}[k_{diff}])}{D}$$

(14)

where arg(.) denotes the argument of a complex variable.

- $z[n]$ is obtained as

$$z[n] = y_c^j[n]c[n]^*$$

(15)

where $y_c^j[n]$ denotes the received PN sequence after the compensation in the $i$th iteration and the residual CFO could be obtained as [17], [18]

$$\hat{\omega}_{\text{fe}} = \frac{\text{arg}(R_{l_{\text{LAG}}})}{l_{\text{LAG}}M}$$

(16)

where $R[l] = \sum_{n=0}^{N_{\text{lag}}-1} z[n]z[n + l]^*$ and $l_{\text{LAG}}$ denotes the lag that the correlation needs.

- The fine estimation of CFO could be obtained by the data auto-correlation $R_{\text{auto}}[k]$, i.e.

$$\hat{\omega}_{\text{fine}} = \frac{\text{arg}(R_{\text{auto}}[k_{\text{auto}}])}{N}$$

(17)

The first one has the largest estimation range which is used in step “frame synchronization” to obtain the coarse estimation of CFO. The first iteration of step “fine estimation of CFO” adopts the second approach to perform more accurate estimation and the following iterations adopt the last one with the highest accuracy to obtain the fine estimation of CFO finally.

2) Iterative channel estimation and equalization: As shown in Fig. 6, we perform coarse estimation and equalization using PN sequence first in this part. Then data-aided channel estimation (DDCE) is used to perform fine estimation and equalization. As described above, the iteration will converge within 3 to 4 iterations, and then we will obtain the data after equalization, as well as CIR and CFR as measurement parameters.

![Block diagram of iterative channel estimation and equalization.](image)

a) Coarse estimation and equalization

The coarse estimation of CIR is obtained by utilizing PN sequence as

$$h_0[n] = \text{IDFT}\left\{\text{DFT}(y_c[n])\right\}$$

(18)

The equalization without reconstruction is performed after channel estimation:

$$\hat{Y}_d(k) = \frac{Y_d(k)}{H^0(k)}$$

(19)

where $Y_d[k]$ and $H^0(k)$ denote the $N_d$-point DFT of $y_d[n]$ and $h_0[n]$, respectively.

b) Hard decision

Since the channel only contains WFS, hard decision is an effective method to eliminate the noise components and recover the transmitted data $d^j[n]$.

c) Data sequence reconstruction

Let $c_j[n]$ and $d_j[n]$ denote FH and FB respectively of the $j$th signal frame. The propagation channel is modeled as a quasi-static $L$th-order FIR filter, where CIR does not vary within the $j$th signal frame period, i.e. the coherence time of channel is much longer than the duration of one signal frame. FH is assumed to be long enough to mitigate the multipath effect ($N_c \geq L$), the received signal will be the linear correlation between the transmitted signal and the CIR as shown in Fig. 7(a), and can be represented as

$$y_{c,j}[n] = c_j[n] * h^j[n], 0 \leq n < N_c + L - 1$$

(20)
\( y_{d,j}[n] = d_j[n] \ast h^i[n], 0 \leq n < N_d + L - 1 \)  \hspace{1cm} (21)

where \( y_{c,j}[n] \) and \( y_{d,j}[n] \) represent the linear convolution of FH and FB with the CIR, respectively. As shown in Fig. 7(b), the shadow areas represent the overlapping in time domain after propagation over a multipath environment of the FH or FB. From (20) and (21), we could obtain the reconstruction circuit convolution of FB in time domain

\[
y_{rd,j}[n] = \begin{cases} 
  y_{d,j}[n] - y_{c,j}[n] + y_{d,j}[n + N_d], & 0 \leq n < L - 1 \\
  y_{rd,j}[n], & L \leq n < N_d - 1
\end{cases}
\]  \hspace{1cm} (22)

which is shown in Fig. 7(c).

d) Channel estimation

The CIR can be obtained as

\[
h^i[n] = \text{IDFT} \left\{ \frac{\text{DFT}(y_{rd,j}[n])}{\text{DFT}(d^i[n])} \right\}
\]  \hspace{1cm} (23)

e) Equalization

The equalization is performed after channel estimation:

\[
\hat{Y}_d(k) = \frac{Y_d(k)}{H^i(k)}
\]  \hspace{1cm} (24)

B. Measurement method for PN595 mode

PN595 is made up of the first 595 symbols of an M-sequence of length 1023, which has no cyclical characteristics. Only different steps from that of performing measurement on PN420 are described in detail in the following.

1) Hard decision: Since PN595 mode usually adopts single-carrier modulation, hard decision is done in time domain.

2) Data sequence reconstruction: This can be done by using the same method as that of PN420 mode, or using the new method named “virtual frame” [19].

Let us define a new transmitted TDS-OFDM block as

\[
s_j = \begin{cases} 
  d_j, & 0 \leq n < N_d \\
  c, & N_d \leq n < N 
\end{cases}
\]  \hspace{1cm} (25)

An extended transmitted TDS-OFDM block is defined as

\[
\bar{s}_j = \begin{cases} 
  c, & 0 \leq n < N_c \\
  s_j, & N_c \leq n < N_c + N 
\end{cases}
\]  \hspace{1cm} (26)

which is shown in Fig. 8. This block includes not only the PN sequence appended in the present block but also the PN sequence appended in the previous block. It’s obvious that the PN sequence at the head of the extended signal block serves as the cyclic prefix, so the reconstruction could be ignored.

3) Equalization: The details of channel estimation and equalization with virtual frame were fully described in [19].

IV. SIMULATION AND TESTING RESULTS

In this section, computer simulation and lab testing are performed to demonstrate the performance of the proposed measurement method, which are all based on the PN420 mode for the reason that other modes have the same performance and precision. The simulation parameters are briefly listed in Table I.

Fig. 9 shows WFS of the modulator. In the simulation, WFS is originated from the non-ideal SRRC filter. Other re-
Fig. 10. WFS of the modulator in the lab testing.

TABLE II
MEASUREMENT RESULTS FROM 20 CONSECUTIVE FRAMES

<table>
<thead>
<tr>
<th></th>
<th>MER(dB)</th>
<th>SFO(ppm)</th>
<th>CFO(Hz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>35.07</td>
<td>16.94</td>
<td>5348.3</td>
</tr>
<tr>
<td>2</td>
<td>34.28</td>
<td>18.91</td>
<td>5351.2</td>
</tr>
<tr>
<td>3</td>
<td>35.84</td>
<td>18.01</td>
<td>5355.2</td>
</tr>
<tr>
<td>4</td>
<td>34.13</td>
<td>21.78</td>
<td>5352.4</td>
</tr>
<tr>
<td>5</td>
<td>35.50</td>
<td>16.98</td>
<td>5352.5</td>
</tr>
<tr>
<td>6</td>
<td>34.82</td>
<td>18.42</td>
<td>5349.4</td>
</tr>
<tr>
<td>7</td>
<td>35.54</td>
<td>17.73</td>
<td>5357.6</td>
</tr>
<tr>
<td>8</td>
<td>35.11</td>
<td>17.16</td>
<td>5348.2</td>
</tr>
<tr>
<td>9</td>
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<td>16.87</td>
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<tr>
<td>10</td>
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<td>16.44</td>
<td>5356.3</td>
</tr>
<tr>
<td>Average</td>
<td>35.15</td>
<td>17.92</td>
<td>5352.3</td>
</tr>
<tr>
<td>Standard deviation</td>
<td>0.59</td>
<td>1.56</td>
<td>3.25</td>
</tr>
</tbody>
</table>

In the lab testing, a real DTMB modulator with PN420 mode is adopted. WFS of the modulator is plotted in Fig. 10. It could be seen that the in-band wave is less than 2dB, however, precision of the evaluation system would be degraded without considering WFS. Table II shows 10 measurement results from 20 consecutive frames. We could clearly see the time-varying property of the modulator.

Fig. 11 shows the measurement results of the same collected data from the original generation measurement system designed by Tsinghua University. The system needs to take a long time to calculate average measurement results with 300 frames. Compared to them, the proposed method adopts only two frames and takes very short time to obtain MER in better precision with more accurate synchronization and channel estimation, and other parameters to reflect the modulator performance in a short time and help users for the modulator troubleshooting.

V. CONCLUSION

This paper presents a novel method to measure the performance of DTMB modulator quickly and accurately. By estimating MER, time/frequency offsets and CIR with only two consecutive frames in an iterative manner, it can precisely reflect the performance of the DTMB modulator in a short time. Both simulation and lab testing demonstrate that our proposed method can evaluate all the key parameters based on the measurement results and reflect the performance of the DTMB modulator precisely. The proposed method is expected to be adopted in the measurement of the DTMB modulator.

REFERENCES


