PIPELINED SIGNAL ADAPTIVE ARCHITECTURE OF A TIME-FREQUENCY WIENER FILTER FOR HIGHLY NONSTATIONARY FM SIGNALS ESTIMATION

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Abstract: Pipelined signal adaptive architecture of an optimal (Wiener) time-frequency (TF) filter has been designed. It is developed based on the real-time results of TF analysis, on the correspondence of the filter's region of support to the instantaneous frequency (IF) of the filtered signal, and on the TF analysis-based IF estimation. The implemented pipelining technique allows the proposed design to overlap in execution unconditional steps performing in neighboring TF instants. In this way, the considered filter significantly enhances critical design performance related to the time required for execution. The achieved improvement corresponds to the one clock cycle by a TF point, which means that the improvement by a single TF point can reach up to 50%. The design is tested on multicomponent highly nonstationary FM noisy signals.

1. INTRODUCTION

Efficient processing of nonstationary signals requires time-varying approaches that can be defined by using common time-frequency distributions (TFDs). Classical TF filters, related to the Richaczek distribution, [1], short-time Fourier (STFT), [1]-[3], and Gabor transform, [4], as well as the Wigner distribution (WD), [5]-[8], exhibit serious drawbacks that limit their applicability. Their extended versions, suppress drawbacks of classical solutions, but they are numerically quite complex and require significant time for calculation. These facts often make the extended solutions unsuitable for real-time analysis. Hardware implementations, when possible, can overcome these problems.

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Single-clock-cycle implementations of TF filters, [1], [2], [9], are quite complex and require repeating of basic calculation elements if they need to be used more than once. Their complexity strongly depends on the estimated signal duration, so they are capable of filtering of signals with the predefined duration only. By considering drawbacks of these solutions, the multiple-clock-cycle signal adaptive hardware implementation of a TF Wiener filter capable of performing real-time estimation and suitable for on a chip implementation has been developed in [11]. It simultaneously improves both critical design performances corresponding to the hardware complexity and performances related to the time required for execution, qualifying itself as an optimal solution in many practical applications. Here we additionally improve the design from [11] and its execution time by introducing the pipelining technique in its development.

The paper is organized as follow. After the introduction, theoretical background is given and the instantaneous frequency (IF) estimation procedure is presented in Section 2. The pipelined implementation of the signal adaptive TF Wiener filter is considered in Section 3. Before the conclusion, in Section 4 the considered pipelined design is tested and verified through the estimation of the multicomponent highly nonstationary FM signal masked by high additive white noise.

2. THEORETICAL BACKGROUND

In frequency domain, optimal (Wiener) nonlinear TF filter can be defined by, [7], [8]:

$$(Hx)(n) = \sum_{k=-N/2+1}^{N/2} L_H(n,k) STFT_x(n,k) \tag{1}$$

where Weyl symbol, $L_H(n,k)$, represents filter's region of support (FRS), [1], [5]-[9], $STFT_x(n,k)=DFT_m[w(m)x(n+m)]$ is the STFT of the q-component noisy signal $x(n)$,

$$x(n) = \sum_{i=1}^{q} (f_i(n)) + \varepsilon(n),$$

$DFT_m[]$ is the discrete FT in m, $w(m)$ is real-valued lag window, and N is the signal duration.

Observe the case of FM signals $f_i(n)$, $i=1,\ldots,q$, highly concentrated in the TF plane around their IFs, and of the additive, widely spread white noise $\varepsilon(n)$, not correlated with the observed FM signals. Following the procedure for the stationary Wiener filter design, [10], in the observed noisy signal case, the FRS of the optimal TF filter corresponds to the combination of IFs of signals $f_i(n)$ [8], [11]. Then, the optimal filtering of nonstationary FM signals can be reduced to the IF estimation in a noisy environment.

In TF analysis framework, the IF estimation is performed by determining frequency points k_i, $i=1,\ldots,q$, where TFD of the noisy signal has local maximum, [8], [11]-[13],

$$IF_i(\vec{n}) = \arg[\max_{k \in Q_{k_i}} TFD_x(n,k)]. \tag{2}$$

In (2), Q_{k_i} is the basic frequency region in TF plane around $f_i(n)$, the IF of which is $IF_i(n)$.

Among all quadratic TFDs, classical WD produces the best IF estimation characteristics in the highly nonstationary monocomponent signals case, [12]. However, based on the full frequency range convolution of input STFT elements (STFTs), used in the classical WD definition, [14]-[17], the WD also produces emphatic cross-terms in the multicomponent signals case, which disables IF estimation in this case and by using classical WD. From the other hand, the cross-terms-free WD (CTFWD) is defined based on the limited convolution of input STFTs (the same STFTs used in the filtering definition (1)),

$$CTFWD_x(n,k) = \left|STFT_x(n,k)\right|^2 + 2\sum_{i=1}^{L(n,k)} \mathrm{Re}\{STFT_x(n,k+i)STFT_x^*(n,k-i)\}. \qquad (3)$$

In (3), $L(n,k) \le L_m$ is the signal adaptive width of the rectangular convolution window, introduced to limit convolution of the STFTs. L_m is its maximum width, determined by the widest STFT auto-term. By definition, in different TF points (n,k), the CTFWD (3) includes variable number of summation terms (the only necessary ones regarding the total energy of each WD auto-term separately). Only the first summation term, $|STFT_x(n,k)|^2 = SPEC_x(n,k)$, is included in calculation (3) outside the STFT auto-terms' domains, their higher number is included inside these domains, whereas the maximum number of these terms, corresponding to L_m, is included only in the central points of the widest STFT auto-term(s). This results

(1) in the CTFWD reduction to the spectrogram (SPEC) outside STFT auto-terms' domains and to the WD inside them,
(2) in the WD cross-terms reduction, or their complete elimination in the non-overlapping multicomponent signals case, [15],
(3) in the noise influence suppression, [15], [17],
(4) in optimization of the IF estimation characteristics among all quadratic TFDs, [12], [13].

In this way, the CTFWD retains the desired characteristic of the WD (including the optimal auto-terms representation, as well as the best IF estimation characteristics) in the monocomponent signals case. However, it also reduces (or, in the non-overlapping multicomponent signals case, completely eliminates) cross-terms of the WD, as well as their influence to the TF signal representation characteristics. Hence, in the non-overlapping multicomponent signals case, IF estimation characteristics of the CTFWD, obtained for each signal's component separately, remain the same as for the case of only that particular component exists, [13].

Moreover, the CTFWD has already been implemented in real-time, [14], and, can be used as an appropriate base in the optimal nonstationary TF filter (1) development, as already performed in [11]. In this paper, the design from [11] is additionally improved by the pipelining technique application.

A. The IF Estimation Procedure [11] Overview

The IF estimation procedure, proposed in [11], and used here in development of the pipelined filter design tests the existence of a local IF in point (n,k) as follows. Frequency-only-dependent CTFWD samples, symmetrically distributed around frequency point k, $CTFWD_x(n,k-L_Q)$, …, $CTFWD_x(n,k)$, …, $CTFWD_x(n,k+L_Q)$, are firstly grouped into, for example, one-dimensional vector Q, sized $2L_Q+1$. In this way, basic frequency interval Q_k (see eq.(2)) is created and the local IF estimation is enabled. The local IF is detected in the frequency point corresponding to the maximum vector element, but only if the maximum vector element is:

(i) the central vector element, $CTFWD_x(n,k)$,
(ii) greater than the introduced spectral floor R,
and if
(iii) the vector size ($2L_Q+1$) satisfies:

$$2\times\max_{1\le i\le q}\{A_i\}\le 2L_Q+1<2\times\min_{\substack{1\le i,j\le q\\ i\ne j}}|IF_i(n)-IF_j(n)| \quad (4)$$

where A_i, $i=1,2,\ldots,q$ are different widths of the non-overlapping CTFWD auto-terms.

The condition (i) in combination with the first inequality from (4), has to be met in order to ensure that:

- All frequency points from the observed auto-term, including the true IF, have the corresponding CTFWD samples inside the vector Q when the existence of the IF in each of these points is investigated. This makes the IF estimation error to be noise-only-dependent inside the CTFWD auto-terms' domains and excludes, from the detection, near-by frequency points around the true IF;
- For each auto-term and each time-instant n, only one frequency point can be detected as a local IF. In this way, the influence of the frequency discretization on the IF estimation quality is reduced, as experimentally proven in [8].

On the other hand, the condition (i) in combination with the second inequality from (4), ensures multiple detection of a local IF in the observed increment of time n, that enables the IF estimation in the case of multicomponent signals.

Finally, the condition (ii) has to be met in order to suppress the noise influence outside CTFWD auto-terms' domains.

After the execution in the k-th frequency point, this procedure is repeated for the frequency point $k+1$. The next frequency-only-dependent CTFWD sample from the same increment of time n, $CTFWD_x(n,k+L_Q+1)$, is imported to the vector Q area, shifting to the left the existing vector elements for one position, [11]. In this way, vector Q slides for one position right over frequency-only-dependent CTFWDs in the observed increment of time, resulting in creating the next basic frequency interval Q_{k+1}. After that, the existence of a local IF in the ($k+1$)-st frequency point is tested in the described manner. For the observed increment of time n, this procedure is applied, point-by-point, to each frequency point k, $k=-N/2+1,\ldots,N/2$.

The presented procedure assumes estimation in the non-overlapping signal components case and allows the IF estimation as long as $\max_{1\le i\le q}\{A_i\}<\min_{1\le i,j\le q,i\ne j}|IF_i(n)-IF_j(n)|$ (i.e. as long as the CTFWD auto-terms' domains occupy comparable widths with the minimal distance between different components).

Finally, parameters L_Q and R of the estimation procedure have to be set. Wide frequency range (4), obtained in the case of highly concentrated, non-overlapping FM signals, suggests the robustness of the IF estimation with respect to the L_Q parameter, proven in [11]. Therefore, the sliding vector Q size of several locations should usually be sufficient. On the other hand, greater values of the spectral floor R almost remove the noise influence outside the CTFWD auto-terms' domains, but they can simultaneously produce significant edge cutting of the finite duration auto-terms (such as chirp signals' auto-terms, considered

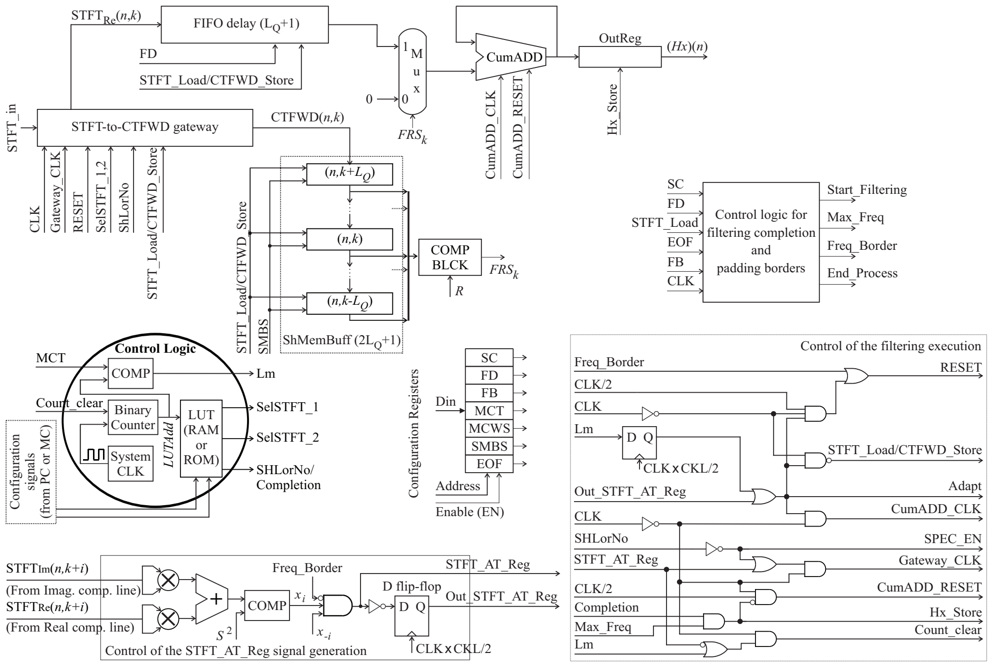

Fig.1. Pipelined hardware implementation of the TF optimal filter based on the CTFWD-related IF estimation.

in this paper, see Section 4). Therefore, the parameter R selection has to be done by making a compromise in the case of finite duration signals. Based on the extensive experimental work, the R value best suited to most applications has been set at 5%-20% of the maximum CTFWD value.

3. PIPELINNED IMPLEMENTATION

Complete pipelined hardware design of an optimal TF filter, principally following definition (1) and based on the CTFWD hardware design, [14], and on the real-time CTFWD-based IF estimation, [11], is given in Fig.1. Modifying input STFT data (STFT_IN signal in Fig.1), the STFT-to-CTFWD gateway, [14], implements the CTFWD calculation (3), producing an improved TF representation of the estimated noisy signal. By each *STFT_Load/CTFWD_Store* cycle, CTFWDs, calculated in the STFT-to-CTFWD gateway, are stored into the ShMemBuff. The ShMemBuff is used to move through the CTFWDs, to produce basic frequency region Q_k, and to take a part in real-time IF estimation procedure implementation (through the implementation of the sliding vector operation). The set of 2-input comparators, comprising in the COMP BLCK, test conditions (i)-(ii) of the IF estimation procedure, developed in [11] and presented in subsection 2.A. If the conditions are satisfied, the COMP BLCK detects FRS in frequency point k, k=–N/2+1,…,N/2, for the observed time instant n, determined by FRS_k=1. The FIFO delay

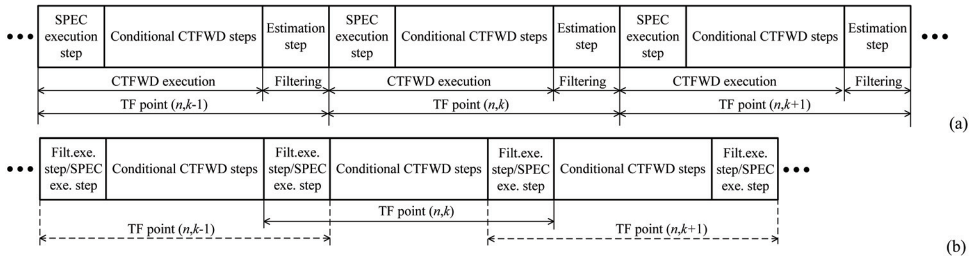

Fig.2. Pipe stages in the TF filter implementation. (a) Non-pipelined implementation from [11], (b) Proposed pipelined implementation.

is used to hold input STFT data and to provide output sample that corresponds in frequency to the ShMemBuff central element.

The design performs the estimation in $L(n,k)+2$ steps per a frequency point, where each of these steps is executed in the corresponding CLK of the filtering execution. In the first $L(n,k)+1$ steps (0-th, 1-st, …, $L(n,k)$-th one) and based on the definition (3), the CTFWD sample is calculated in the STFT-to-CTFWD gateway, [14]. In each of these steps, the corresponding summation term from (3) is produced, taking part (or not) in the CTFWD calculation. Only summation terms existing inside STFT auto-term's domains take part in the calculation (3). The first detected summation term, non-existing inside STFT auto-terms' domains, is not included in summation (3). However, it also terminates the CTFWD calculation in the observed TF point. The TF filter function is then implemented in the ($L(n,k)+1$)-st–estimation–step, which is overlapped in execution by the 0-th step (SPEC execution step) of the next frequency point $k+1$.

Only the SPEC execution step and the estimation one are unconditional. These steps provide the SPEC-based IF estimation. Residual steps are conditional and depend on the estimated signal shape. They are used to improve the IF estimation quality up to the CTFWD-based one and are taken only in TF points existing inside the STFT auto-terms' domains, determined by the signal adaptive period of the *STFT_AT_Reg* signal. Unity value of the *STFT_AT_Reg* signal detects summation term that should be included in calculation (3), while its zero value implies following actions:

(1) Through the participation in the *Gateway_CLK* signal generation, disables the i-th term ($i=1,\dots,L(n,k)$) to enter the CTFWD calculation (3) in i-th CLK, and
(2) Through the participation in the *STFT_Load/CTFWD_Store*, *RESET* and *CumADD_CLK* signals generation, terminates (in the same CLK) the summation (3).

For the observed TF point, this results in completion of the CTFWD calculation, but also in the TF filtering execution in the following CLK (next conditional (($i+1$)-st) one). In this way, the *STFT_AT_Reg* signal allows the proposed design both

(1) to optimize the number of CLKs taken in different TF points within the execution and
(2) to produce the CTFWD-based IF estimation.

The *STFT_AT_Reg* signal also controls the filtering completion in the observed frequency point k, $k=-N/2+1,\dots,N/2$.

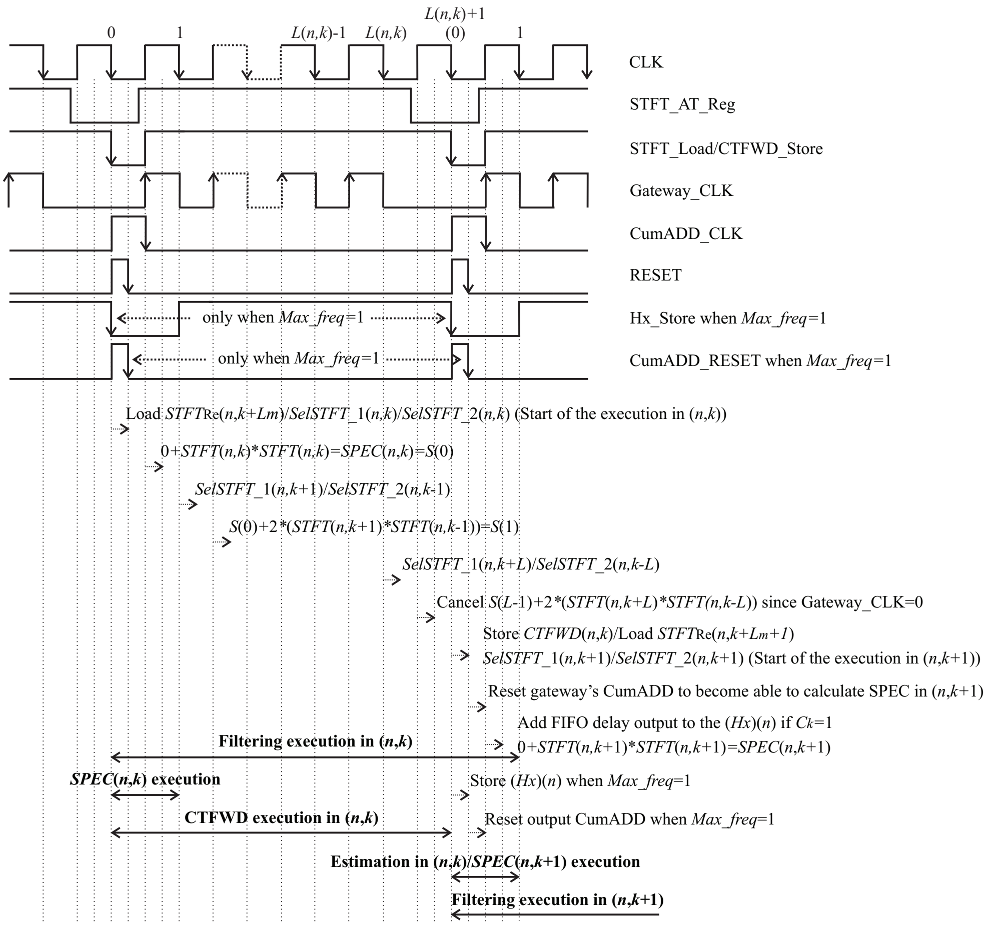

Fig.3. Timing diagram of signals that control execution in the proposed pipelined implementation.

Following the development from [11], in our design the final–completion–step of a time instant n is performed after the execution in each frequency point from n and is overlapped in execution with the SPEC execution step from the next TF point $(n+1,-N/2+1)$. In this way, the improvement in execution time by only one CLK per a time instant n is achieved.

The proposed pipelined design additionally allows overlapping in execution of the unconditional steps between the substantial frequency points k, $k+1$, $k=-N/2+1,\ldots,N/2$, Fig.2, improving the execution time by one CLK, but per a frequency point. This can be a significant development in comparison to the hardware design from [11] because each time instant contains N frequency points. Residual steps cannot be included in pipelining, because they are conditional and do not have to exist. Signals *STFT_Load/CTFWD_Store, CumADD_CLK*, *Gateway_CLK*, *RESET,* and *CumADD_RESET* control the calculation in the STFT-to-CTFWD gateway, the summation in the CumADD in a frequency point k,

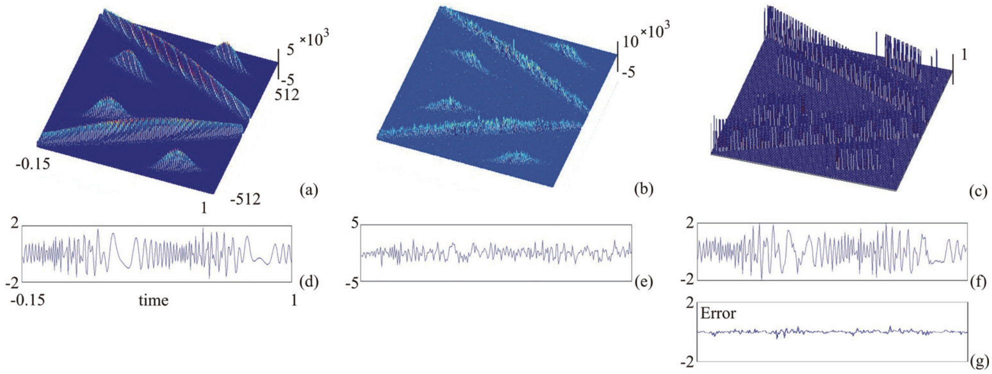

Fig.4. (a) CTFWD of the non-noisy signal *f*(*t*); (b) CTFWD of the noisy signal; (c) Estimated IF/FRS; (d) Signal *f*(*t*); (e) Noisy signal; (f) Output signal of the proposed pipelined architecture, implemented in FPGA, (g) Filtering error.

$k=-N/2+1,\ldots,N/2$, the filtering completion in a time instant n, as well as the pipelining, as shown Fig.3. Generation of these signals, shown in Fig. 1, slightly increases hardware complexity, but also decreases capacity of the look-up-table memory required by the implementation.

By using pipelining technique, the proposed design improves throughput of the implementation that corresponds to a CLK per a TF point. In comparison to the design developed in [11] and depending on the sliding vector size L_Q and on the normalized signal rate, the improvement can reach values of about 15% (for L_Q=7) in frequency points existing around IFs, up to the 50% in frequency points existing outside STFT auto-terms' domains.

4. TESTING AND VERIFICATION

Verification of the proposed pipelined design has been performed through filtering of the 3-component test signal

$$f(t)=e^{-45(t-2/25)^2}\cos(900(t+1.3)^2)+e^{-(t-2/5)^2}\times$$
$$\times\cos(1200(t+0.3)^2)+e^{-45(t-2/3)^2}\cos(900(t-1/22)^2) \qquad (5)$$

considered within the time interval [–0.15,1], Fig.4(d). This signal is masked by high white noise, Fig.4(e), such that input signal-to-noise ratio (SNRin) takes value SNR_{in}=–0.34[dB]. Parameters T_w=0.25, $R=0.05\max_{n,k}\{CTFWD_x(n,k)\}$, N=256, L_Q=5, and reference level applied in the CTFWD calculation, [14]-[16], of $0.1\cdot\max_{n,k}\{|STFT_x(n,k)|^2\}$ are used. The design verification has been performed through the implementation in the Stratix II family EP1S10F780C5 device.

In practice, signals usually spread across a range of frequencies whose width grows with nonstationarity of their characteristics. In the case of linear FM signals, considered in this example, the range of frequencies occupied by each signal component separately dominantly depends on the normalized rate signal of this components (grows with the

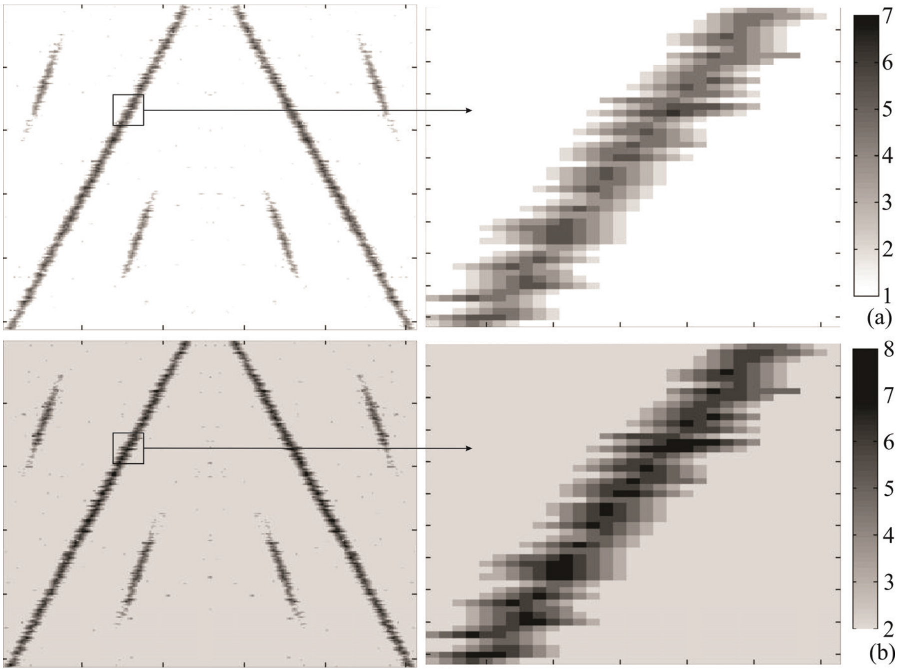

Fig.5. Distribution of CLKs performed per a TF point. (a) Pipelined implementation, (b) Non-pipelined implementation.

normalized signal rate growing). In that sense, the signal (5) can be considered as the highly nonstationary, since normalized signal rates of its components are 0.828, 0.844, 0.828, respectively.

Results of real-time estimation, performed in the noisy signal (5) case, are represented in Fig.4. Output SNR of 17.03[dB] and the SNR improvement of 17.37[dB] have been achieved. Knowing that (5) is highly nonstationary signal, the achieved improvement can be considered as very high. Besides, maximum improvement of up to approximately $(156/N)\times10\log(N/4)+(100/N)\times10\log(N/2)=19.24$[dB] is expected, but only theoretically, in a partly 2-component (in 100 time instants) and a partly 4-component (in 156 time instants) signal case.

To prove and to visually represent the achieved improvement, distribution of CLKs taken by the proposed pipelined design per a frequency point in the noisy signal (5) case is shown in Fig.5(a), and is compared with the non-pipelined design from [11], Fig.5(b). For the observed case, the improvement can easily be noted, computed, and numerically expressed by 45.358%.

To verify logic of the proposed pipelined implementation, the FPGA simulation (on the arbitrary numerical values, but on the developed design and the control logic) is performed and results are given in Fig.6. The improvement of a CLK cycle per a TF point can easily be observed. Since TF points existing outside the STFT auto-terms' domains are considered, the highest possible improvement in execution time of 50% (in these points) is

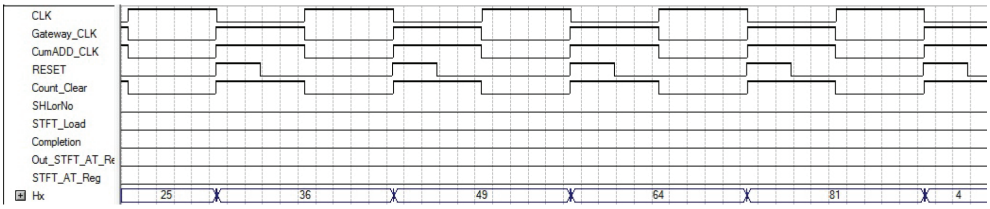

Fig.6. Verification of the proposed pipelined implementation performed in FPGA.

represented in Fig.6. These observations are in full correspondence with the theoretical consideration about the improvement in execution time (of a CLK cycle per a TF point), achieved by the proposed pipelined design. These can be easily noted by consideration the *Hx* signal from Fig.6, whose numerical value duration per a TF point corresponds to the duration of one CLK cycle.

5. CONCLUSIONS

The proposed hardware design retains desirable characteristics of multiple-clock-cycle signal adaptive designs from [11] and [14], regarding calculation and implementation complexity, as well as the execution time. Moreover, by applying the pipelining technique, the design additionally improves TF filtering execution time approximately up to the about 45%, depending on the estimated signal shape. In this way, it overcomes the corresponding TFD-based filters regarding almost all critical design criteria. Besides, it enables high quality real-time TF filtering, based on the highest quality signal adaptive CTFWD-related IF estimation. Non-adaptive designs and stationary filters cannot produce so high quality results.

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