DSP IMPLEMENTATION OF SQRT(M)-BEST ROTATED QAM SOFT-DEMAPPER

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ABSTRACT

In the second generation terrestrial digital video broadcasting (DVB-T2) systems, one of the important features is the usage of rotated quadrature amplitude modulation (QAM) to exploit signal space diversity. However, the implementation of soft-demapper is very challenging and becomes more complex as the modulation order increases. Recently, a $\sqrt{M}$-best soft-demapper was proposed which can achieve the same performance as the full search scheme but with low complexity. In this paper, we implement the $\sqrt{M}$-best soft-demapper on a digital signal processer (DSP) with a coarse-grained reconfigurable array (CGRA) architecture. The best candidate selection algorithm is accordingly modified for friendly implementation. Our results show that the $\sqrt{M}$-best soft-demapper can be implemented with 82% less cycles than the full search based soft-demapper for the most complex 256-QAM case.

Index Terms— DVB-T2, rotated QAM, soft-demapper, $\sqrt{M}$-best, DSP implementation

1. INTRODUCTION

The second generation terrestrial digital video broadcasting (DVB-T2) standard has been developed for the advanced digital terrestrial transmission system offering higher efficiency, robustness and flexibility [1]. It adopts the latest modulation and coding techniques to increase the spectrum efficiency, e.g., modulation up to 256 quadrature amplitude modulation (QAM), orthogonal frequency division multiplexing (OFDM) of up to 32768 points, and low-density parity-check (LDPC) coding with code word lengths of up to 64800 bit, which thus poses critical real-time requirements on the implementation of the receivers. In addition, the receivers are required to be flexible to allow the variations of the modulation scheme, the number of OFDM subcarriers, the pilot pattern and so on [2]. Therefore, it is recommendable to base the receiver designs on the software-defined radio (SDR) paradigm [3]. In [4, 5], the software implementation of the DVB-T2 receiver has been studied in the state-of-the-art digital signal processors (DSPs) and graphics processing units (GPUs).

In DVB-T2, the rotated QAM constellation is introduced to improve performance, especially for very frequency selective channels. This comes at the expense of increase complexity of soft-demapper when compared to the non-rotated QAM case. In the DVB-T2 implementation guidelines [2], the Max-Log approximation based full search soft-demapper is recommended, in which the log-likelihood ratio (LLR) is obtained by calculating the Euclidian distances from the received signal to all the constellation points, and then finding the corresponding minimum ones for each bit. However, many studies have shown that the implementation of the Max-Log full search soft-demapper is too complex for the high modulation order cases [4–6].

To reduce the complexity of soft-demapper, one possible approach is to reduce the number of constellation candidates. In [7] and [8], the full constellation is decomposed into four overlapped subregions (or subsets). According to the sign of received signal, only the points in one subregion are selected for LLR calculation. In [9], a geometrical approach was proposed, which tried to directly find the best candidate without searching. With the aid of geometrical transformations, it first determined the closest lines to the received signal for each bit. Then, the closest point in the closest line was considered as the one with minimum distance. Even though the complexity is somehow reduced in [7–9], all these works can not achieve the same performance as the Max-Log full search scheme.

Recently, the authors proposed a $\sqrt{M}$-best soft-demapper, which only selects $\sqrt{M}$ best candidates for LLR calculation of either even or odd bits, but achieves the same performance as the full search scheme using M constellation points [10]. In this paper, we implement the $\sqrt{M}$-best soft-demapper on a DSP core based on a coarse-grained reconfigurable array (CGRA) architecture. To enable friendly implementation, the algorithm of the best candidate selection part is redesigned. With appropriate constellation transformation, the best candidate selection can become much easier than performing it in the originally faded constellation. Furthermore, the fixed difference between the coordinates of the best candidates is also utilized for complexity reduction. Our results show that the $\sqrt{M}$-best soft-demapper is suitable for DSP implementation and the vector processing feature can be fully utilized. For 64-QAM and 256-QAM, its cycles are respectively 63% and 82% lower than that of the full search scheme.

2. BACKGROUND

We consider a bit-interleaved coded modulation (BICM) system with rotated QAM transmissions. In the system, n coded

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and interleaved bits $\{b_0, b_1, \cdots, b_{n-1}\}$ are mapped on a M-QAM symbol $s = s_I + js_Q$ ($M = 2^n$), where s_I conveys the even bits $\{b_0, b_2, \cdots, b_{n-2}\}$, and s_Q conveys the odd bits $\{b_1, b_3, \cdots, b_{n-1}\}$ [1]. With a rotation angle α, the rotated QAM symbol can be written as

$$\begin{aligned}\tilde{s} &= e^{j\alpha}s = \tilde{s}_I + j\tilde{s}_Q \\ &= (s_I \cos\alpha - s_Q \sin\alpha) + j(s_I \sin\alpha + s_Q \cos\alpha). \end{aligned} \quad (1)$$

To further obtain diversity, the Q part is cyclically delayed by one cell (subcarrier) within a forward error correction (FEC) block. After cell interleaving, the I and Q parts which belong to the same symbol are transmitted on two different cells with independent fading.

At the receiver, the OFDM demodulation and phase compensation are performed in each cell. After that, the I and Q parts which belong to the same symbol are collected from the corresponding cells again. The received signal is equivalently written as

$$r = r_I + jr_Q = h_I\tilde{s}_I + jh_Q\tilde{s}_Q + w \quad (2)$$

where h_I and h_Q denote the amplitudes of the fading channel on the two cells, and w is the noise with variance σ^2.

To perform soft-demapping, we need to calculate the LLR of r for each bit b_i, which is defined as

$$LLR(b_i) = \frac{1}{\sigma^2}[\min_{\tilde{s}\in\mathbb{S}^0_{b_i}} D(\tilde{s}) - \min_{\tilde{s}\in\mathbb{S}^1_{b_i}} D(\tilde{s})], \quad (3)$$

where $\mathbb{S}^1_{b_i}$ and $\mathbb{S}^0_{b_i}$ are the sets of rotated constellation points with b_i being 1 and 0, respectively, and

$$D(\tilde{s}) = (r_I - h_I\tilde{s}_I)^2 + (r_Q - h_Q\tilde{s}_Q)^2 \quad (4)$$

is the Euclidean distance between the received signal r and the constellation points $\tilde{s}$ (after fading). Note that the Max-Log approximation is applied to obtain the LLR form in (3). The straightforward approach is to calculate the Euclidean distances between the received signal and all M constellation points, and then find the minimum ones for each bit case [2]. However, the complexity of this Max-Log based full search scheme is too high for implementation in the real systems, especially for the high modulation order cases.

3. $\sqrt{M}$-BEST SOFT-DEMAPPER

In [10], the authors proposed to simplify the soft demapping by utilizing the constellation mapping feature. For any even bit b_i, the points in the same column always belong to the same set ($\mathbb{S}^0_{b_i}$ or $\mathbb{S}^1_{b_i}$). While for the odd bit, the points in the same row always belong to the same set. After the constellation is rotated and faded, this mapping feature is still valid. Thus, in each row and column, it is enough to select only the best candidate (the one with minimum distance to the received signal) for distance comparison with other candidates, instead of selecting all the points. In other words, for either even or odd bits, we only need to select $\sqrt{M}$-best candidates for distance comparison, and it is as valid as the full search case in terms of finding the point with minimum distance. [1]

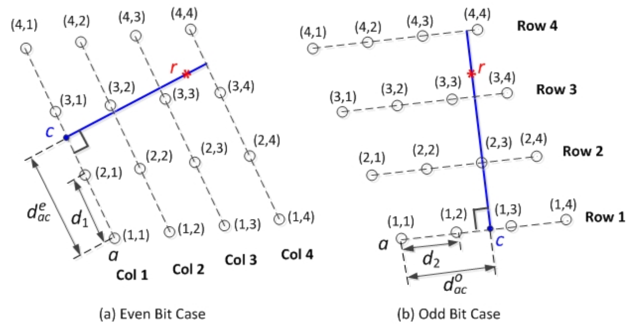

Fig. 1: Best candidate selection for the even and odd bits.

The $\sqrt{M}$-best soft-demapper is performed in the following steps:

1. Select $\sqrt{M}$-best candidates for the even bits, and $\sqrt{M}$-best candidates for the odd bits;
2. Calculate the Euclidean distances between the received signal and $2\sqrt{M}$ best candidates;
3. Calculate the LLR for each bit.

In Step 1, the method to find the best candidate is projecting the received signal to each column and row, and then obtaining the coordinates of the cross point. In each column and row, the candidate closest to the cross point is selected as the best one, as shown in Fig. 1. This may involve calculating the distance between a reference point to the cross point, and quantization processing.

3.1. Implementation-Friendly Best Candidate Selection

In this part, we design an implementation-friendly algorithm for the best candidate selection part. The basic idea is to find the best candidate in alternative constellation, instead of the faded constellation. This will not affect the final results if the received signal and projection line are also accordingly transformed.

An example of the faded constellation and received signal is depicted in Fig. 2(a), and the projection line which aims to find best candidates for even bit case is also shown. If the fading impact in the received signal is compensated, the signal is transformed to the original rotated QAM constellation, as shown in Fig. 2(b). If further anti-rotated, it is transformed to the original non-rotated QAM constellation, as shown in Fig. 2(c). Since the projection line is also accordingly transformed, it may not be perpendicular to the columns anymore.

[1]Even though the constellation is rotated and faded, we keep using row and column to describe the corresponding lines as in the original constellation (as shown in Fig. 1).

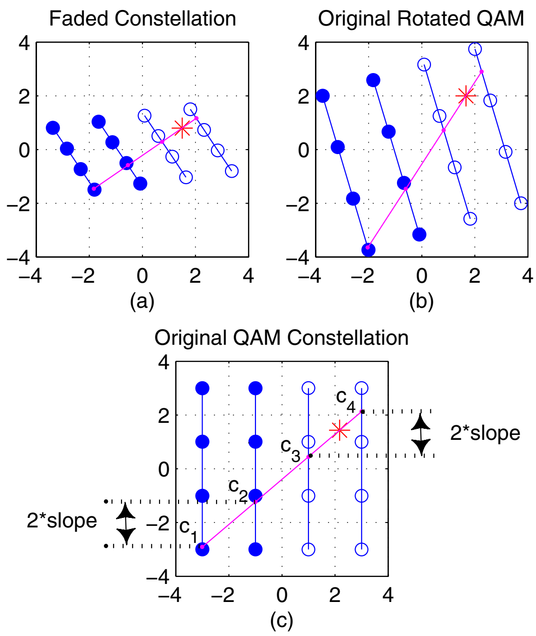

Fig. 2: An example of best candidate selection in the original QAM constellation for the even bit case.

Instead, it is just a sloped line passing through the transformed signal. The slope of the projection line in three constellation cases in Fig. 2 is $\frac{h_I \sin\alpha}{h_Q \cos\alpha}$, $\frac{h_I^2 \sin\alpha}{h_Q^2 \cos\alpha}$ and $\frac{(h_I^2 - h_Q^2)\sin\alpha\cos\alpha}{h_Q^2 \cos^2\alpha + h_I^2 \sin^2\alpha}$, respectively. However, the relative positions of the cross points in each constellation case keep same because the transform is linear. Therefore, we propose to calculate the coordinates of cross points in the original non-rotated QAM constellation, because one part (I/Q) of the cross points is known. In addition, the quantization processing to find best candidates is also easier because the interval of neighbor candidates is 2, and it is suitable for implementation.

Furthermore, the fixed difference between the coordinates of the neighbour cross points can be utilized to simplify the calculation. In the original QAM constellation, the difference of the I part is 2, and the Q part 2ρ, where ρ is the slope of the projection line, as shown in Fig. 2(c). After calculating the coordinate of the first cross point, the others can be easily obtained by adding the fixed difference. By applying the above features, the best candidate selection algorithm is redesigned in Algorithm 1.

Algorithm 1 Implementation-Friendly Best Candidate Selection

Assume the input is equalized signal $\hat{s}$ (the fading impact is compensated from the received signal);
Calculate the transformed signal $t = \hat{s}e^{-j\alpha} = t_I + jt_Q$, $\rho_e = \frac{(h_I^2 - h_Q^2)\sin\alpha\cos\alpha}{h_Q^2\cos^2\alpha + h_I^2\sin^2\alpha}$, $\rho_o = \frac{(h_I^2 - h_Q^2)\sin\alpha\cos\alpha}{h_I^2\cos^2\alpha + h_Q^2\sin^2\alpha}$;
Calculate $a_I = a_Q = -\sqrt{M} + 1$;
for $i = 1 : \sqrt{M}$ **do**
  % For Even Bits Case
  Calculate $d = (t_Q - a_Q) - \rho_e(t_I - a_I) + (i-1)(2\rho_e)$;
  Find the index of best candidate, $k = round(\frac{d}{2} + 1)$; if $k < 1$, $k = 1$; if $k > \sqrt{M}$, $k = \sqrt{M}$;
  Save index $K_i^e = (k, i)$;
  % For Odd Bits Case
  Calculate $d = (t_I - a_I) - \rho_o(t_Q - a_Q) + (i-1)(2\rho_o)$;
  Find the index of best candidate, $k = round(\frac{d}{2} + 1)$; if $k < 1$, $k = 1$; if $k > \sqrt{M}$, $k = \sqrt{M}$;
  Save index $K_i^o = (i, k)$.
end for

4. DSP IMPLEMENTATIONS AND RESULTS

4.1. DSP Architecture

The processor considered for implementation is an enhanced version of the one considered in [4]. As shown in Fig. 3, it has a architecture that constitutes combination of a very long instruction word (VLIW) processor and a coarse-grained array (CGA) accelerator. It has totally 12 function units (FUs) which consist of three scalar FUs, three vector pack/unpack FUs, two vector load/store FUs, one 32-bit vector arithmetic and three 512-bit vector arithmetic FUs. The 12 FUs can work simultaneously for the CGA accelerator. Among three scalar FUs, two are shared between the VLIW processor and CGA accelerator. The 32-bit vector FU supports 2-way (with 16-bit elements) single instruction multiple data (SIMD) processing and 512-bit vector FU supports 32-way (with 16-bit elements) SIMD processing and 16-way (with 32-bit elements) SIMD processing.

Two exclusive operating modes are available, i.e., VLIW mode and CGA mode. Massive data processing algorithms that are iterative are mapped to the CGA mode, while the general control instructions and simple sequential data processing codes are mapped to the VLIW mode. In VLIW mode, instructions are fetched from VLIW instruction cache, decoded and executed on VLIW FUs. In CGA mode, instructions are loaded from the CGA configuration memory and executed concurrently in all of 12 FUs. In this mode, the instruction-level parallelism (ILP) is maximized with the efficient software pipelining and proper scheduling of the interconnection among FUs.

4.2. Implementation Results

We verify the implementation results of four soft-demapper kernels, including Max-Log based full search [2], sub-region [7], geometrical [9], and $\sqrt{M}$-best. The FEC block has a size of 64800 bits, with different combinations of QAM level and number of cells (16-QAM with 16200 cells, 64-QAM with 10800 cells, or 256-QAM with 8100 cells). The fixed-point

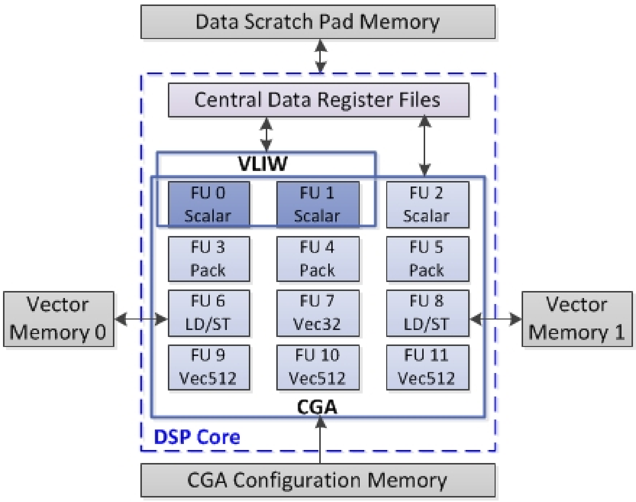

Fig. 3: DSP Architecture.

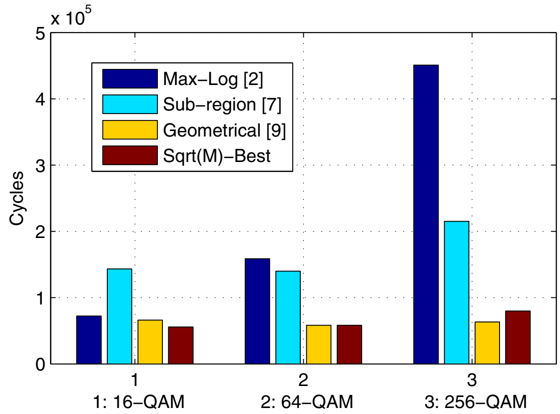

Fig. 4: Complexity comparison of different kernels in terms of number of required cycles.

data input to the soft-demapper has 16-bit precision, i.e., a complex sample has 32 bits (with 16-bit real part and 16-bit imaginary part). The LLR output has 8-bit precision.

The $\sqrt{M}$-best soft-demapper kernel can fully utilize 512-bit SIMD processing, which supports to process 16 cells in a parallel way. Its implementation is straightforward and composed of three CGA loops, 1) Q-delay processing, 2) Best candidate selection and Euclidean distance calculation, and 3) LLR calculation. The full search kernel is implemented in a similar way, but it calculates the Euclidean distance for all the candidates in the 2nd CGA loop. In the sub-region soft-demapper kernel, direct SIMD processing is difficult because it requires to check the sign of the signal and select different candidate set for different cell. Therefore, the total cells are first separated into 4 groups according to the sign of the signal. The SIMD processing can then be utilized in each group. For the implementation of the geometrical soft-demapper kernel, the SIMD processing can also be directly utilized, but a vector version of lookup table is required to find the closest constellation lines parallelly.

In Fig. 4, the kernel mapping results are shown in terms of required cycles when performing soft-demapper for one FEC block. Compared to the full search kernel, the required number of cycles in the $\sqrt{M}$-best kernel is 23%, 63%, and 82% lower, respectively for 16-QAM, 64-QAM and 256-QAM cases. The sub-region kernel has much higher cycles than the $\sqrt{M}$-best one due to the extra grouping cost. Since the grouping overhead becomes higher as the number of cells increases, the number of total cycles in 16-QAM and 64-QAM is comparable. The cycles of the geometrical soft-demapper kernel are similar for different QAM cases, because the number of dominated instructions is linear to the number of transmitted bits. Only in the 256-QAM case, it has lower cycles than the $\sqrt{M}$-best kernel.

The soft-dempper kernels are plugged into the full chain of the DVB-T2 receiver to verify the performance. Except for the soft-dempper kernel, the other kernels are almost similar as the ones considered in [4], but modified based on the current processor architecture. The simulations are performed in the memoryless Rayleigh fading channel with various combination of modulation order and code rate. When the $\sqrt{M}$-best soft-demapper kernel is used, the bit error rate (BER) performance is same as that with Max-Log based full search kernel. However, the subregion and geometrical kernels always have performance loss when compared to the full search case. The signal-to-noise-ratio (SNR) gaps at 10^{-4} BER in different simulation cases are listed in Table 1. It is observed that the SNR gap becomes larger in the lower modulation order and less robust code rate.

Table 1: SNR Gap (dB) of Soft-Demapper Kernels Compared to Full Search Kernel in Memoryless Rayleigh Channels (at 10^{-4} BER)

QAM Order	16	16	64	64	256	256
Code Rate	3/5	4/5	3/5	4/5	3/5	4/5
$\sqrt{M}$-Best	0	0	0	0	0	0
Sub-region [7]	0.15	0.19	0.08	0.14	0.01	0.03
Geometrical [9]	0.23	0.98	0.17	0.37	0.05	0.13

5. CONCLUSIONS

In this paper, we have implemented and compared several soft-demappers for rotated QAM on a CGRA DSP. The $\sqrt{M}$-best soft-demapper can fully utilize the vector processing features and requires less cycles in most situations, while the other schemes need special pre-processing or intrinsics for vector processing. The number of cycles for most complex 256-QAM using $\sqrt{M}$-best soft-demapper is similar as that for the 16-QAM using full search kernel. In addition, there is no performance loss when compared to the full search kernel.

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