On the Impact of Fixed Point DSP Implementation on Required Channel Estimator Complexity in Communication Receivers

Asghar Havashki #1, Per Gunnar Kjeldsberg #2, Geir E. Øien #3, Lars Lundheim #4, Jan T. Nymoen #5

Department of Electronics and Telecommunications
Norwegian University of Science and Technology
NO-7491, Trondheim, Norway
{ [1]havashki, [2]Per.Gunnar.Kjeldsberg, [3]Geir.Oien, [4]lundheim, [5]nymoen }@iet.ntnu.no

***Abstract*— In this paper we discuss the impact that fixed point implementation of DSP units in a communication system chain has on the required complexity of channel estimators. We have restricted our model to a flat-fading channel and linear FIR estimators, under the assumption that finite word length effects in the DSP operations in the communication chain can be modeled as additional additive noise. We have shown that a small increase in this noise can lead to a considerable increase in the required estimator complexity if a given NMSE performance for the channel estimation must be upheld, in particular at medium-to-high CSNRs.**

I. Introduction

In wireless communications, the power consumption associated with the actual transmission has traditionally been assumed to be dominant in the overall power budget. However, taking digital signal processing (DSP) and RF circuitry power consumption into account becomes increasingly relevant as communication distance decreases [1]. Therefore, co-optimization of circuitry and transmission schemes, and energy efficient DSP design, has received much attention recently.

Orthogonal Frequency Division Multiplexing (OFDM) is one example of a technique which in recent years has become widely applied in wireless communication systems, due to its high data rate transmission capability and high bandwidth efficiency [2]. However, the performance of OFDM and other spectrally efficient schemes depends, among other things, on advanced digital signal processing (DSP) and on the use of efficient and possibly adaptive resource allocation and transmission techniques. These in turn require that *accurate estimates of the channel* are available in the receiver and transmitter. The importance of accurate channel estimation is also demonstrated in [3], a study on the impact of channel estimation error on the performance of linear FIR equalizers. It shows that as $\frac{E_s}{N_0}$ approaches 20 dB and beyond, the influence of the channel estimation error on the overall error-rate becomes important for the scenario discussed. For lower $\frac{E_s}{N_0}$, the additive channel noise dominates the overall error-rate. In this paper we are concerned with wireless communication scenarios where the system performance is primarily limited by available estimation accuracy.

Accurate channel estimation of a time and frequency dispersive wireless fading channel calls for complex estimators, which might lead to significant energy dissipation in such devices. While a lot of research is done on analysis and optimization of different estimation techniques in various system contexts ([4], [5], [6]), our aim in this paper is rather to study the impact that *fixed point implementation of DSP units* in the communication chain has on the required complexity for a certain chosen class of estimators. We do not claim that this class is necessarily the best choice for all systems. It is rather chosen due to its simple (FIR filter) structure and analytical tractability.

Since in DSP realizations data must be represented by a finite number of bits, designers are forced to make trade-offs between precision and complexity. As power has become a major constraint in design of e.g., ad hoc wireless networks, saving power by reducing the number of bits when realizing DSP operations is an attractive approach. The corresponding reduction in the level of precision of the DSP operations can often be modeled as introducing *additional additive noise* to the signal being operated on. This is also the approach used here. Such a noise model can easily be applied to, e.g., a simple scalar quantizer [7]. Examples of how the modeling can be done for more advanced DSP operations like FFT/IFFT can be found in [8][9], and for other important OFDM receiver components in [10]. We will in this paper simply assume that this model holds, without detailing particular DSP operations, and then study the impact of such additional noise on the required complexity of the channel estimator if a certain overall system performance (according to some chosen performance metrics) is to be upheld. The chosen measure of estimator complexity is the FIR filter order used to realize the estimator.

Trade-off situations related to these issues might occur if designers wish to reduce the number of bits when realizing a particular DSP unit, in order to reduce energy dissipation in the corresponding unit. The consequence of such an action would then be a certain increase in the additional noise introduced to the signal. This increase in noise subsequently needs to be compensated for, e.g., by increasing the channel estimator order, if the goal is to uphold the estimator performance. This in turn comes with an increase in energy dissipation in the channel estimator.

Results from our study can help designers to make correct choices in such situations.

II. System Model

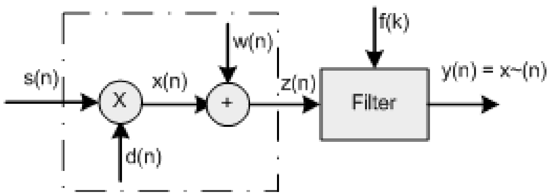

Fig. 1. System model

In our system model (Figure 1) a stream of complex-valued symbols, $s(n)$, with pilot symbols equally spaced within the input stream with period L, enters a time varying Wide Sense Stationary (WSS) flat fading channel. This might model, e.g., one sub-channel in an OFDM system. The channel estimates are derived from only the pilot symbols, which for simplicity are all assumed to be of equal power, with an absolute value of 1. In the channel, the symbols are subjected to fading, with $d(n)$ denoting the complex fading gain. A complex-valued Additive White Gaussian Noise (AWGN) $w(n)$ is then added to the faded signal $x(n)$, so that the resulting noisy signal $z(n)$ entering the channel estimator becomes:

$$z(n) = s(n) \cdot d(n) + w(n). \tag{1}$$

In (1) $w(n)$ is assumed to be the total noise added to the pilot symbols from all relevant sources, including the additional noise resulting from the assumed fixed point realization of DSP operations in the system, denoted $w_{\text{d}}(n)$. Thus

$$w(n) = w_{\text{c}}(n) + w_{\text{d}}(n) \tag{2}$$

where $w_{\text{c}}(n)$ is the ambient noise, while $w_{\text{d}}(n)$ will depend on the DSP implementation used. Our aim is now to see how $w_{\text{d}}(n)$ affects the estimator's performance and subsequently how the estimator complexity must be adapted in order to maintain a certain performance level. The actual value of $w_{\text{d}}(n)$ will depend on the particular DSP operations included in the communication chain, and the details of implementation of these operations. A specific example of $w_{\text{d}}(n)$ can be the additional noise introduced to the signal due to fixed point realization of an FFT unit in the receiver, when the number of bits used to represent the *data path* and the *twiddle factors* are limited. Table I shows the error variance in *mean square error* (MSE), contributed from one FFT operation [11], due to fixed point realization of such a unit, for different FFT sizes. The error variance is calculated as the deviation between the output of floating point and fixed point realizations of the FFT, averaged over the number of operations, i.e.:

$$\text{MSE} = \frac{1}{M} \sum_{i=0}^{M-1} \mid y(i) - \hat{y}(i) \mid^2 \tag{3}$$

with $y(i)$ and $\hat{y}(i)$ being the outputs of the floating point and fixed point FFT, respectively, and M being the number of operations. The simulations are based on 1000 operations ($M = 1000$) with Gaussian input. The number of bits used in the data path was for simplicity chosen to be equal to the number of bits used to represent the so-called twiddle factors.

III. Channel Estimator

The task of the channel estimator is to provide an estimate of $d(n)$ in (1) which is good enough for the overall communication system to work according to specifications, such as requirements on the average *bit error rate* (BER). As mentioned in the introduction, we shall assume that a linear FIR estimator is to be used; it is well known from estimation theory that this is MMSE optimal if the channel exhibits Rayleigh fading. To estimate the channel gain $d(n)$, we employ a non-causal lowpass filter which uses a number of the received pilot symbols, both from the past and future:

$$\hat{d}(n) = \sum_{k=-N/2}^{N/2} f(k) z(n-k) \tag{4}$$

The *optimal* filter coefficient vector in the *maximum a posteriori* (MAP) or minimum MSE sense, $\mathbf{f}_{\text{MAP}} = [f(\frac{-N}{2}), ..., f(\frac{N}{2})]_{\text{MAP}}$, for such an estimator of order N, is given by [12]:

$$\mathbf{f}_{\text{MAP}}^T = \mathbf{r}^T (\mathbf{R} + \frac{\mathbf{1}}{\bar{\gamma}} \mathbf{I})^{-\mathbf{1}} \tag{5}$$

where $\mathbf{I}$ is the $N \times N$ identity matrix, $\mathbf{R}$ is a normalized covariance matrix (a symmetric Toeplitz matrix), $\mathbf{r}$ is a normalized covariance vector, and $\bar{\gamma}$ is the average Channel Signal-to-Noise Ratio (CSNR). By assuming *isotropic scattering* for our fading environment the *Jakes* fading model can be applied [12], which simplifies the computation of $\mathbf{r}$ and $\mathbf{R}$. We have used the Jakes model in our simulations, but this is not a critical choice for our work.

The channel estimation error $e(n)$, is now computed as:

$$e(n) = d(n) - \hat{d}(n) \tag{6}$$

Typically, a certain *normalized mean square error* (NMSE), NMSE $\triangleq \frac{E[|e(n)|^2]}{E[|d(n)|^2]}$, is required if a specific system performance is to be guaranteed. This means that we must ensure NMSE $\leq \triangle_{\text{max}}$, where the value of $\triangle_{\text{max}}$ depends on the type of system at hand and on the desired performance level. For example, in [13], a study on adaptive trellis-coded modulation (TCM) over predicted flat fading channels, it is demonstrated how NMSE of the channel *prediction*[1] error influences the *average spectral efficiency*. It shows that, for any given average CSNR of 20 dB or above, the average spectral efficiency decreases noticeably if the NMSE increases from 10^{-2} to 10^{-1}. This impact is considerably larger as the target *bit error rate* (TBER) decreases, e.g., from 10^{-3} to 10^{-5}.

[1] The difference between our considered estimation and the prediction performed in [13] is simply that we consider estimation of the current channel gain, whereas the system in [13] explicitly predicts future channel gains.

Word Length	FFT-16	FFT-64	FFT-256	FFT-1024	FFT-4096
18	$6.25 \cdot 10^{-11}$	$2.18 \cdot 10^{-10}$	$3.24 \cdot 10^{-10}$	$3.90 \cdot 10^{-10}$	$4.88 \cdot 10^{-10}$
16	$1.87 \cdot 10^{-9}$	$3.12 \cdot 10^{-9}$	$5.07 \cdot 10^{-9}$	$6.25 \cdot 10^{-9}$	$8.54 \cdot 10^{-9}$
14	$3.43 \cdot 10^{-8}$	$5.78 \cdot 10^{-8}$	$7.81 \cdot 10^{-8}$	$9.76 \cdot 10^{-8}$	$1.36 \cdot 10^{-7}$
12	$5.37 \cdot 10^{-7}$	$8.75 \cdot 10^{-7}$	$1.36 \cdot 10^{-6}$	$1.66 \cdot 10^{-6}$	$2.17 \cdot 10^{-6}$
10	$8.12 \cdot 10^{-6}$	$1.40 \cdot 10^{-5}$	$2.11 \cdot 10^{-5}$	$2.53 \cdot 10^{-5}$	$3.41 \cdot 10^{-5}$
8	$1.25 \cdot 10^{-4}$	$2.34 \cdot 10^{-4}$	$3.39 \cdot 10^{-4}$	$4.29 \cdot 10^{-4}$	$5.61 \cdot 10^{-4}$
6	$2 \cdot 10^{-3}$	$3.9 \cdot 10^{-3}$	$5.8 \cdot 10^{-3}$	$1.04 \cdot 10^{-2}$	$3.07 \cdot 10^{-2}$

TABLE I

RESULTING ERROR VARIANCE (MSE) DUE TO FIXED POINT REALIZATION OF FFT OF DIFFERENT SIZES, COMPARED TO A FLOATING POINT REALIZATION, WHEN DIFFRENT WORD LENGTH IS USED.

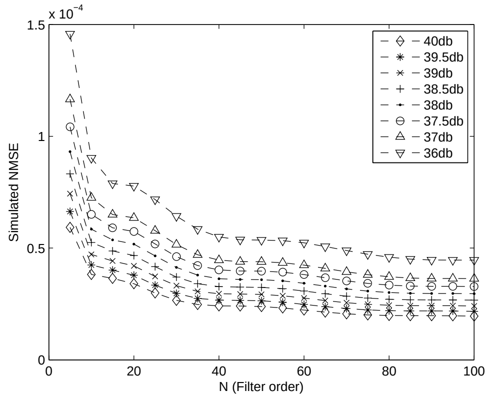

Fig. 2. Simulated NMSE as function of N for CSNR of 40, 39.5, 39, 38.5, 38, 37.5, 37 and 36 dB. Doppler rate equals 0.005, and results are for 1000000 simulations.

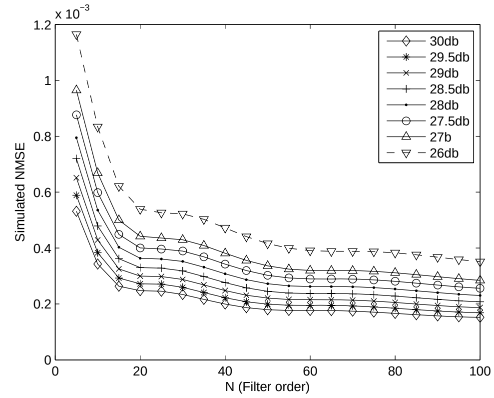

Fig. 3. Simulated NMSE as function of N for CSNR of 30, 29.5, 29, 28.5, 38, 27.5, 27 and 26 dB. Doppler rate equals 0.005, and results are for 1000000 simulations.

The study further shows that for this particular application, this influence is small when the NMSE is between 10^{-3} and 10^{-2}. For other purposes however, such as for coherent receiver detection in various systems, even stricter demands must be placed on channel estimation NMSE [14].

Typically, the actual value of the NMSE will depend on the CSNR, i.e. implicitly on the variance of $w(n)$ for a given transmit power, as well as on estimator order (longer filter yields lower NMSE), and channel correlation. This means that there will be a trade-off between the accuracy of the DSP implementation and the accuracy of the estimator. If we allow an increase in $w_{\text{d}}(n)$, and thus in $w(n)$, fewer bits are needed in the DSP implementation. This will, however, decrease the performance for a given estimator order.

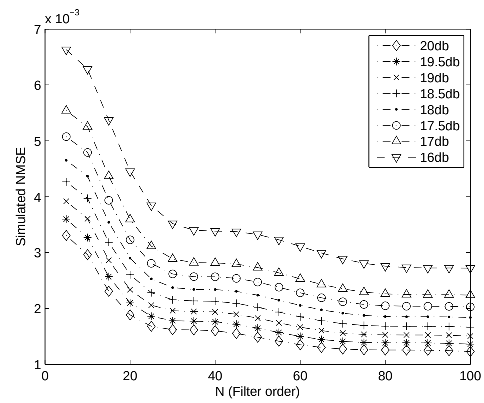

Fig. 4. Simulated NMSE as function of N for CSNR of 20, 19.5, 19, 18.5, 18, 17.5, 17 and 16 dB. Doppler rate equals 0.005, and results are for 1000000 simulations.

IV. RESULTS AND DISCUSSION

We have modeled our system using the SystemC language and design environment [15]. This will enable us to perform fast and accurate exploration also of the more hardware related implementation details in our future studies of this problem. We have run simulations for different estimator filter orders (N) and CSNR levels (10 dB, 20 dB, 30 dB and 40 dB), where the CSNR is supposed to take into account both ambient noise and DSP implementation noise. For each case we have reduced the CSNR by 0.5 dB, 1 dB, 1.5 dB, 2 dB, 2.5 dB, 3 dB and 4 dB (corresponding to fewer and fewer bits being used in DSP implementation, while keeping the ambient noise constant) to see how such a reduction in CSNR affects the required estimator order for a given performance, measured in normalized estimation error variance. The results show that even a CSNR reduction of 0.5 dB leads to a considerable increase in the required estimator order (N) in all cases, if the goal is to keep the NMSE constant. Figure 2 shows the NMSE as function of N when the initial CSNR (corresponding to infinite word lengths in the DSP implementation) is 40 dB.

We can divide Figure 2 into three regions. In the first region, NMSE falls steeply with increasing N and is reduced from its maximum value to 42% of its maximum. For a CNSR

Initial Level	region 1	Initial N at transition	Added N in region 1	region 2	N at transition	Added N in region 2	Added N in region 3
40dB	100 - 42%	8	1	42 - 27%	28	4-9	>>18
30dB	100 - 28%	13	2-3	28 - 19%	39	6-15	>>25
20dB	100 - 18%	21	1-8	18 - 14%	54	10-22	>>30
10dB	100 - 12%	33	1-11	12 - 10%	73	17-32	>>50

TABLE II

ASSUMED REQUIRED NMSE LEVELS WITH NEEDED N AND ADDITIONAL NEEDED N AS CSNR IS REDUCED BY 0.5 DB.

Initial Level	region 1	Initial N at transition	Added N in region 1	region 2	N at transition	Added N in region 2	Added N in region 3
40dB	100 - 47%	7	2	47 - 30 %	24	8-10	>>25
30dB	100 - 32%	11	2-4	32 - 21%	34	7-21	>>35
20dB	100 - 20%	17	2-7	20 - 14 %	48	10-32	>>50
10dB	100 - 13%	29	2-15	13 - 11%	40	32-57	>>60

TABLE III

ASSUMED REQUIRED NMSE LEVELS WITH NEEDED N AND ADDITIONAL NEEDED N AS CSNR IS REDUCED BY 1 DB.

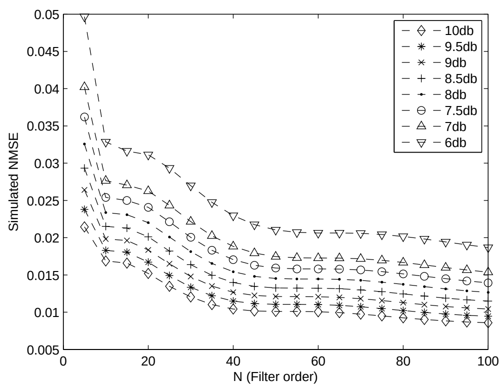

Fig. 5. Simulated NMSE as function of N for CSNR of 10, 9.5, 9, 8.5, 8, 7.5, 7 and 6 dB. Doppler rate equals 0.005, and results are for 1000000 simulations.

reduction of 0.5 dB, e.g., from 40 dB to 39.5 dB, N must be increased only by 1 in order to maintain the same performance. For the 40 dB case, $N = 8$ at the 42% value. In the second region, NMSE decreases more slowly with increasing N, and lies between 42% and 26% of its maximum value. In this region, a CSNR reduction of 0.5 dB requires that N is increased by 4 to 9 to maintain the performance. From this point on, in the third region, NMSE decreases slowly as N increases, and N must be increased very much (>>18) to keep up the NMSE performance for an extra 0.5 dB reduction in CSNR. A similar behavior can be seen when the initial CSNR is 30 dB, 20 dB and 10 dB (Figures 3, 4 and 5, respectively) instead of 40 dB. The percentage values defining the transitions between the three regions are different, however, as summarized in Table II. Tables III, IV, and V summarize results for the cases where the CSNR is reduced by 1 dB, 2 dB and 3 dB, respectively.

We will now see how the noise contributed from one DSP operation (due to fixed point realization of such a unit) might impact the required complexity of the estimator. We use the quantization error resulting from fixed point realization of an FFT unit, as given in Table I [11], as a practical example.

Table I shows the resulting added noise due to fixed point realization, compared to a floating point realization, for FFTs with different sizes (16, 64, 256, 1024, 4096). Each row gives the results for different word lengths (number of bits used in the DSP implementation). The same word length is assumed used for both the data path and twiddle factors. The resulting noise for FFT-16 and 10 bits word length is about $8.12 \cdot 10^{-6}$. This corresponds to about 0.35 dB reduction in CSNR when the initial CSNR is 40 dB, and about 0.035 dB reduction in CSNR when the initial CSNR is 30 dB. The effect of the corresponding noise on the channel estimation order is consequently negligible even for these high CSNR conditions.

If we reduce the number of bits to 8 when realizing the FFT-16, the resulting noise will be on the order of $1.25 \cdot 10^{-4}$ (Table I). This is equal to 3.52 dB and 0.51 dB reduction in CSNR when the initial CSNR is 40 dB and 30 dB, respectively. Looking at the results in Table V (or Figure 2) we find that when the CSNR is 40 dB, the NMSE is reduced to 68% of its maximum value when the estimator order approaches 4. To uphold the same performance when the CSNR is reduced by 3 dB, an estimator order increase of 5-7 is required. To further reduce the NMSE to about 47% of its maximum, the estimator order must be increased to 7 when the CSNR is at its initial value of 40 dB. To achieve the same reduction in NMSE for the case where CSNR is reduced by 3 dB, the estimator order must be increased by 20-29. Using Table II (or Figure 3) we furthermore find that when the CSNR is at an initial value of 30 dB, an NMSE reduction to 28% of its maximum value is achieved when the estimator order approaches 13. An additional estimator order of 2-3 is required to achieve a performance at the same level when the CSNR is reduced by 0.5 dB. Further reduction of NMSE to 19% of its maximum value is achieved, for the initial case (30 dB), as the estimator order approaches 39. To achieve the same performance when CSNR is reduced by 0.5 dB, an additional estimator order of 6-15 is required. A noise increase of $1.25 \cdot 10^{-4}$ has only minor effect on the estimator order when the initial CSNR is 20 dB or below.

If the number of bits in the FFT-16 module is further reduced to 6, it causes an additional noise contribution of $2 \cdot 10^{-3}$ (Table I), which equals a CSNR reduction of 13.2

Initial Level	region 1	Initial N at transition	Added N in region 1	region 2	N at transition	Added N in region 2	Added N in region 3
40dB	100 - 59%	5	3-5	59 - 37%	24	14-51	>>60
30dB	100 - 37%	9	5-9	37 - 27%	14	23-37	>>50
20dB	100 - 24%	14	2-14	24 - 19%	19	31-53	>>85
10dB	100 - 15%	22	2-27	%			>>50

TABLE IV

ASSUMED REQUIRED NMSE LEVELS WITH NEEDED N AND ADDITIONAL NEEDED N AS CSNR IS REDUCED BY 2 DB.

Initial Level	region 1	Initial N at transition	Added N in region 1	region 2	N at transition	Added N in region 2	Added N in region 3
40dB	100 - 68%	4	5-7	68 - 47%	7	20-29	>>50
30dB	100 - 44%	7	3-14	44 - 32%	11	28-50	>>90
20dB	100 - 32%	8	2-16	32 - 28%	11	19-24	>>40
10dB	100 - 36%	3	1-4	36 - 20%	7	17-37	>>60

TABLE V

ASSUMED REQUIRED NMSE LEVELS WITH NEEDED N AND ADDITIONAL NEEDED N AS CSNR IS REDUCED BY 3 DB.

dB, 4.8 dB, 0.8 dB, and 0.08 dB when the initial CSNR is at 40 dB, 30 dB, 20 dB, and 10 dB, respectively. A noise contribution of such an order (13.2 dB reduction in CSNR) clearly destroys the estimation results for the CSNR level of 40 dB. For the case where the initial CSNR is at 30 dB, a CSNR reduction of about 4.8 dB calls for dramatic increase in the estimator order if the goal is to uphold the performance of the initial case (Figure 3). When the initial CSNR is 20 dB, the NMSE is reduced to 18% and 14% of its maximum value when the estimator order approaches 21 and 54, respectively. An additional increase of 1-8 and 10-22 in the estimator order is required to reduce the NMSE to the same level as for the initial cases. The effect of the corresponding noise for the case where CSNR is at 10 dB is negligible.

This example clearly demonstrates the fact that as the number of bits to realize an FFT of size 16 is reduced, the system-level effects of additional noise introduced to the signal increase. The increase in noise must be compensated for by an increase in the estimator order if the goal is to uphold the NMSE performance. The amount of increase in noise also depends on the size of the FFT transform. As FFT size is increased, the additional noise generated using the same number of bits is higher (Table I), requiring higher estimator order. The following example shows how estimation results and estimator order will be affected if an FFT of size 64 is used instead of 16.

In this case, when the number of bits to represent data and twiddle factors is chosen to be 10, the resulting additional noise is about $1.4 \cdot 10^{-5}$. This is equal to a CSNR reduction of 0.57 dB and 0.06 dB when the initial CSNR is 40 dB and 30 dB, respectively. Results summarized in Table II show that in order to reduce the NMSE to 42% of its maximum value for the initial case of 40 dB, the estimator order required is 8. When the CSNR is reduced by about 0.5 dB, the channel estimator order must be increased by 1 in order to achieve the same performance in this region. Further improvement by reducing the NMSE to 27% of its maximum value is achieved for the initial case when the estimator order approaches 28. An increase of 4-9 in estimator order is required to achieve the same performance when the CSNR is reduced by about 0.5 dB. As the word length is further reduced to 8 bits, an additional noise variance of $2.34 \cdot 10^{-4}$ is produced. This is equal to a 5.24 dB, 0.91 dB, and 0.10 dB reduction in CSNR when the initial CSNR is 40 dB, 30 dB and 20 dB, respectively. A reduction of CSNR by 5.24 dB when the initial value is 40 dB, destroys the estimation results. For a CSNR of 30 dB, the NMSE is reduced to 32% of its maximum value when the estimator order is increased to 11. The estimator order must be increased by about 2-4 to achieve a performance at the same level in this region, when the CSNR is reduced by about 1 dB. Further improvement by reducing the NMSE to 21% of its maximum value is achieved for the initial case when the estimator order approaches 34. An increase of 7-21 in estimator order is required to achieve the same performance when the CSNR is reduced by about 1 dB. The corresponding noise variance of $2.34 \cdot 10^{-4}$ has minor effect on the channel estimator order for the case where the CSNR is 20 dB and below. Reducing further the word length to 6 bits generates a noise variance of about $3.9 \cdot 10^{-3}$. This is equal to 16 dB, 6.9 dB, 1.4 dB and 0.16 dB when the initial value is 40 dB, 30 dB, 20 dB and 0.6 dB, respectively. As shown in the above examples, the impact of the produced additional noise on the estimator order is insignificant for all CSNR levels when a 10 bits word length is used for an FFT-16. For an FFT-64, a 10 bits word length has impact on the required estimator order when the initial CSNR is 40 dB. Use of 8 bits word length causes a significantly larger increase in estimator order when an FFT-64 is used compared to using an FFT-16. This trend holds also for larger FFT sizes.

The results presented above can be used by designers while implementing communication systems. As an example let us assume that the designer knows that the system will be mainly used for a certain channel condition, e.g., CSNR = 30 dB, and with a given estimation accuracy requirement, e.g., NMSE $= 0.25 \cdot 10^{-3}$. Using the methodology presented in this paper, it is possible to perform a trade-off between the number of bits in the fixed-point DSP implementation, and the channel estimator complexity. If, for simplicity, we assume that the only DSP unit involved is an FFT-16 module, the results from Figure 3 and Table I can be used. With an FFT-16 module implemented with 10 bits, the designer find that the impact of using fixed point instead of floating point is negligible. If, to reduce FFT power consumption, the designer wants to decrease its number of bits to 8, the increase in error variance is $1.25 \cdot 10^{-4}$, equal to a CSNR reduction of 0.5 dB. From Figure 3 the designer can see that this will require an increase in estimator filter order of 14 to uphold the estimation accuracy. Similar trade-off scenarios can be depicted for other channel conditions, estimation accuracy requirements, and DSP modules. Such systematic analysis has the potential to greatly simplify the design work.

V. Conclusions

In this paper we have discussed the impact that fixed point implementation of DSP units in the communication system chain has on the required complexity of channel estimators. We have restricted our model to a flat-fading channel and linear FIR estimators, under the assumption that finite word length effects in the DSP operations in the communication chain can be modeled as additional additive noise. We have shown that a small increase in this noise can lead to a considerable increase in the required estimator complexity if a given NMSE performance for the channel estimation must be upheld, in particular at medium-to-high CSNRs. As part of future work, we will take finite-word-length effects in the estimator itself into account. We will also, by using results for relevant computations simulated in *Synopsys Power Compiler*, translate complexity into energy dissipation and show how the mentioned variations in word length and estimator order changes the overall energy dissipation. We will also target a more analytical characterization of trends that so far have been observed based on simulation.

References

[1] A. Goldsmith, *Wireless Communications*, 1st ed. Cambridge University Press, 2005.

[2] X. Wang and H. V. Poor, *Wireless Communication Systems: Advanced Techniques for Signal Reception*, 1st ed. Prentice Hall, 2004.

[3] D. Welden, F. Simoens, H. Steendam, and M. Moeneclaey, "Impact of channel estimation errors on the performance of linear FIR equalizers for frequency selective MIMO channels," in *Proc. SCVT'05*, Enschede, The Netherlands, Nov. 2005.

[4] D. Lowe and X. Huang, "Adaptive low-complexity MMSE channel estimation for OFDM," in *Proc. ISCIT*, Bangkok, Thailand, Oct. 2006.

[5] E. Lohan and M. Renfors, "Novel adaptive filter for fading channel estimation in coherent CDMA receivers," in *Proc. ISCCSP*, March 2004.

[6] J.Beek, O. Edfors, M. Sandell, P. Börjesson, and S. Wilson, "On channel estimation in OFDM systems." in *Proc. VTC*, Chicago, IL, USA, July 1995.

[7] J. G. Proakis and D. G. Manolakis, *Digital Signal Processing, Principles Algorithms and Applications*, 4th ed. Prentice Hall, 2007.

[8] L. Lundheim, "Influence of DFT/FFT coefficient quantizing on OFDM receiver performance," in *Proc. Norwegian Symposium on Signal Processing*, Bergen, Norway, Oct. 2003.

[9] R. Perlow and T. Denk, "Finite wordlength design for VLSI FFT processors," in *Proc. ACSSC*, Pacific Grove, CA, USA, Nov. 2001.

[10] M. Speth, S. A. Fechtel, G. Fock, and H. Meyr, "Optimum receiver design for wireless broad-band systems using OFDM - part i," *IEEE Transactions on Communications*, vol. 47, no. 11, pp. 1668–1677, Nov. 1999.

[11] J. T. Nymoen, "Enhanced studies of the fixed point FFT," Norwegian University of Science and Technology, Contact: nymoen@iet.ntnu.no, Tech. Rep.

[12] H. Holm, "Adaptive coded modulation performance and channel estimation tools for flat fading channels," Ph.D. dissertation, Norwegian University of Science and Technology, 2002, available at: www.henrikholm.com/avhandling/thesis.pdf.

[13] S. Falahati, A. Svensson, M. Sternad, and H. Mei, "Adaptive trellis-coded modulation over predicted flat fading channels." in *Proc. VTC*, Orlando, Oct. 2003.

[14] H. Meyr, M. Moeneclaey, and S. A. Fechtel, *Digital Communication Receivers Synchronization, Channel Estimation, and Signal Processing*. John Wiley & Sons, 1998, pp. 741–747.

[15] Systemc 2.0.1 language reference manual. [Online]. Available: www.systemc.org.