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COMPLEXITY EVALUATION FOR THE IMPLEMENTATION OF A PRE-FFT EQUALIZER IN AN OFDM RECEIVER

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ABSTRACT

A Pre-FFT Equalizer (PFE) has been shown to offer a significant throughput efficiency improvement when applied to an OFDM receiver. Alternatively, the PFE can be used to increase the maximum delay spread conditions under which the OFDM system can operate effectively. Due to the manner of its operation, the PFE requires the use of modified adaptation algorithms if iterative, decision directed, adaptation is required. In this paper, the computational complexity required to implement a PFE and a suitable adaptation strategy is evaluated. Initially, an LMS adaptation algorithm is investigated and evaluated in terms of its suitability for application in conjunction with the PFE to standards such as ETSI DVB-T and HIPERLAN/2 and IEEE 802.11a. The complexity requirements are found to be high, particularly in the case of DVB-T. The demand for a lower complexity adaptation algorithm is thus identified. As a result, a CSI-based adaptation method is subsequently considered. The complexity requirement of this algorithm is also analyzed and evaluated and is shown to be much lower than that of the LMS algorithm. Thus, it is shown that if the CSI-based adaptation method is used, the dominant complexity requirement is due to the implementation of the equalizing filter and not the adaptation method. Reduced filter complexity requirement is thus shown to be the key to enabling effective application of the PFE. The ATSC 8-VSB standard is identified as a possible source of techniques to reduce or facilitate the high complexity demands for implementation of the PFE filter.

I. INTRODUCTION

Orthogonal Frequency Division Multiplexing (OFDM) is a robust modulation technique that has been selected for a number of radio communications standards, including DVB-T [1], HIPERLAN /2 [2] and IEEE P802.11a [3]. These standards are expected to have significant impact in the consumer electronics market, particularly in areas such as digital video distribution and home wireless networking.

A novel combined OFDM-Equalization technique [4], incorporating a pre-FFT Equalizer (PFE) has recently been developed. This technique has been shown to offer an improvement in bandwidth efficiency over the conventional OFDM technique [5]. Alternatively, the PFE can be used to increase the maximum delay spread duration under which an OFDM system can operate effectively. This results in the potential application of OFDM systems in scenarios beyond those that they were originally designed for. For example, HIPERLAN/2 could be used in larger scale outdoor environments or DVB-T 2k could be used in a single frequency network. Whatever the application of the PFE, the improvement comes at the expense of additional receiver complexity.

The performance of the PFE has already been investigated under radio impairments such as additive noise and mobile channel conditions [5]. A brief summary of the Combined OFDM-Equalization technique and the function of the PFE is given in section II. A more detailed description is given in [5].

In order to be compatible with the function of the PFE, standard iterative adaptation techniques such as the LMS algorithm require a slight modification if they are to be used in a decision directed manner. The required modification is discussed in section III.

This paper investigates the computational complexity required to implement the PFE and a suitable adaptation algorithm. The required number of complex operations for PFE filter implementation is determined in section IV.

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Initially, adaptation using the LMS algorithm is considered since this provides a suitable reference for comparison with other technologies. An analysis of the complexity requirement of this algorithm is given in section V for the cases of training and decision directed adaptation.

The analysis undertaken in sections IV and V is used to determine the additional complexity cost of applying combined OFDM-equalization in comparison to conventional OFDM for each of the three OFDM based standards mentioned above. This additional complexity cost is considered alongside the efficiency gains offered by combined OFDM-equalization in section VI.

Due to the high computational complexity requirements of an LMS adapted PFE (particularly in the case of DVB-T), a lower complexity option is also considered in this paper. In section VII, an adaptation method, based on Channel State Information (CSI) is proposed. This is investigated in terms of computational complexity in section VIII.

The computational complexity of the CSI-based adaptation method is evaluated in section IX.

Conclusions are drawn in section X.

II. COMBINED OFDM-EQUALIZATION & THE PRE-FFT EQUALIZER

A combined OFDM-Equalization receiver incorporating a pre-FFT equalizer is shown in figure 1. This receiver is compatible with a standard OFDM transmitter, provided that regular training sequences are inserted into the transmitted signal for channel estimation and/or equalizer training.

The receiver shown in figure 1 takes as its input the baseband received sequence $y'(n,l)$. n indexes the transmission symbols and l indexes the OFDM symbols. Each OFDM symbol consists of $(N+M)$ transmission symbols of duration T_s. N transmission symbols form the useful symbol period and M transmission symbols form the guard band.

The received sequence is filtered by the pre-FFT equalizer to produce the equalized sequence $z'(n,l)$. The equalizer takes a similar form to a conventional Linear Transverse Equalizer (LTE) or Decision Feedback Equalizer (DFE) and its tap coefficients are termed $c(j,n,l)$. j indexes the equalizer taps and J_1 and J_2 are used to denote the number of taps in the feedforward and feedback (if present) sections of the equalizer. It is assumed that the equalizer is clocked at the transmission symbol rate $1/T_s$.

The guard interval is then extracted and an FFT applied to produce the frequency domain received vector $Z(k,l)$.

The frequency domain vector is used to generate a channel estimate $S(k,l)$. This is commonly referred to as Channel State Information (CSI). The channel estimator typically requires some a-priori knowledge of training sequences or pilot symbols inserted into the transmitted signal $x'(n,l)$.

A channel compensation process, based on the CSI, is subsequently applied to $Z(k,l)$ to produce $V(k,l)$. This is followed by a decision process that generates the output data.

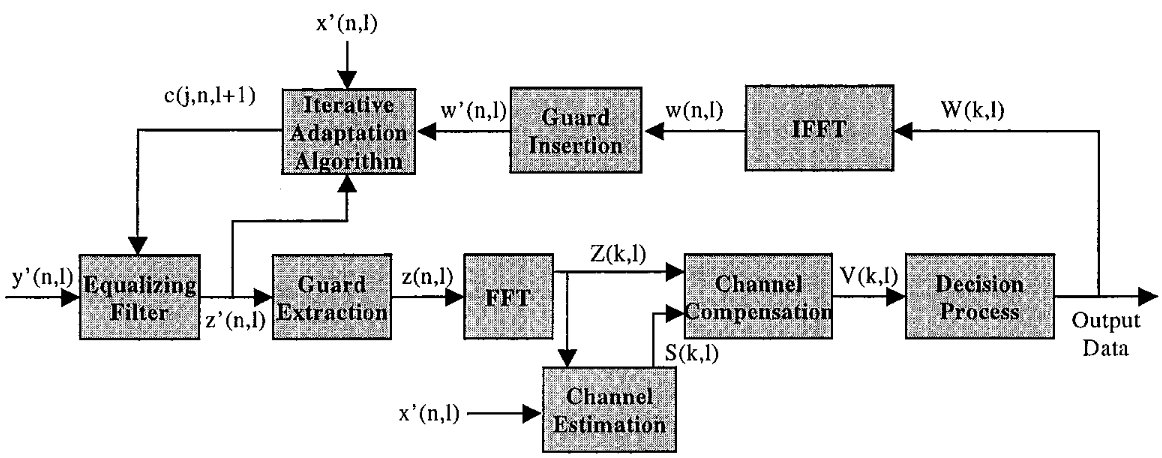

Figure 1. Combined OFDM-Equalization Receiver Supporting Decision Directed Adaptation

If the equalizer is adapted by training then the iterative adaptation algorithm takes the equalizer output $z'(n,l)$ and a reference to the training sequence $x'(n,l)$ as its inputs and generates the equalizer tap coefficient vector for the next transmission symbol period, $c(j,n+1,l)$. Conventional adaptation algorithms can be used for iterative equalizer training.

In order to support decision directed equalizer adaptation, a decision feedback vector $w'(n,l)$ is generated from the post decision data symbols. This decision feedback vector can be used in an iterative adaptation algorithm to generate the equalizer tap coefficient vector for the next OFDM symbol period, $c(j,n,l+1)$. Conventional adaptation algorithms cannot be used for iterative, decision directed, equalizer adaptation. Instead, a modified algorithm of the general form described in section III must be employed.

III. A PFE COMPATIBLE METHOD FOR ITERATIVE, DECISION DIRECTED, ADAPTATION

An iterative equalizer adaptation algorithm can be described in general form by:

$$c(j,n+1)=c(j,n)+f(c(j,n)) \qquad (1)$$

Hence, the tap coefficient vector for the next transmission symbol period is updated according to a function of the current tap vector. For an error based algorithm such as the LMS, the tap coefficients are updated as a function of the error in the equalizer's output. This proves problematic in the case where the PFE is required to be updated in a decision directed manner due to the function of the FFT. The FFT takes N time domain input symbols and generates N frequency domain output symbols at intervals of $N+M$ transmission symbols. Thus, for $N+M-1$ transmission symbols, no estimate of the equalizer output error can be obtained. To accommodate this limitation a modified algorithm is used for iterative adaptation. This takes the general form:

$$c(j,n,l+1)=c(j,n,l)+\sum_{n=0}^{N-1} f(c(j,n,l)) \qquad (2)$$

Thus, after a complete OFDM symbol has been received, the tap coefficient vector for the next OFDM symbol is updated according to a summation of the N updates that would have been applied if an estimate of the equalizer's output error had been available previously. This approach has been shown to support accurate tracking provided that the channel does not change significantly during a single OFDM symbol period [8].

IV. FILTER COMPLEXITY ANALYSIS

The computational requirement of the PFE filter is a function of the total number of taps employed and is identical to the computational requirements of a conventional LTE or DFE.

The number of complex mulitplications required per clock is given by:

$$N_{CMULT-F}=(J_1+J_2) \qquad (3)$$

The number of complex additions required per clock is given by:

$$N_{CADD-F}=(J_1+J_2-1) \qquad (4)$$

V. LMS ANALYSIS

As discussed in sections II and III, the PFE requires the use of a modified adaptation algorithm in order to support decision directed adaptation. A conventional adaptation algorithm can be used for equalizer training. In this section, the computational requirements for the use of the LMS algorithm for both training and decision directed adaptation are investigated.

LMS-Training: The standard LMS algorithm can be described by:

$$c(j,(n+1),l)=c(j,n,l)+\Delta y(n-j)\varepsilon(n,l)^* \qquad (5)$$

$\varepsilon(n,l)$ denotes the equalizer output error.

The required number of complex multiplications per clock cycle assuming LMS Training are given by:

$$N_{CMULT-T}=(J_1+J_2) \qquad (6)$$

The required number of complex additions per clock cycle assuming LMS training is given by:

$$N_{CADD-T}=(J_1+J_2+1) \qquad (7)$$

Note that a further $(J_1 + J_2)$ real multiplications are also required to implement the multiplication by Δ. However, if Δ has a value which is a reciprocal value of 2, these multiplications can be implemented as a simple bit shift [9]. Since this is a very simple operation to implement, these real multiplications are neglected from subsequent analysis.

LMS Decision Directed Tracking: Decision Directed adaptation of the PFE employs a variation on the conventional LMS algorithm taking the general form [5]:

$$c(j,n,(l+1)) = c(j,n,l) + \sum_{n=0}^{N-1} \Delta y(n-j)\varepsilon(n)^* \qquad (8)$$

This modified adaptation algorithm generates one new coefficient vector per OFDM symbol, instead of per transmission symbol. Thus, the coefficient vector must be updated at *1/N* times the symbol rate during training, with each update requiring N times more operations per update. The summation term in equation 8 is implemented by reproducing both the equalizing filter functionality and the LMS-Training algorithm within the decision directed LMS coefficient calculation process. The required computation for decision directed LMS adaptation is thus equal to the sum of the filter and LMS-training computation requirements.

The required number of complex multiplications per clock cycle for LMS decision directed adaptation is:

$$N_{CMULT-DD} = 2(J_1 + J_2) \qquad (9)$$

The required number of complex additions per clock cycle for LMS decision directed adaptation is given by:

$$N_{CADD-DD} = 2(J_1 + J_2) \qquad (10)$$

VI. COMPLEXITY EVALUATION FOR THE LMS ALGORITHM

In order to evaluate the computational complexity requirements for adaptation of the PFE by means of the LMS algorithm, three further steps are required.

Firstly, since an additional IFFT is required to support the generation of the decision feedback vector, the additional complexity this entails must be evaluated. The computational requirement of an FFT or IFFT is given by [7]:

$$N_{CMULT-FFT} = \frac{N \log_2 N}{2} \qquad (11)$$

Secondly, the number of complex multiplications and additions must be translated into real operations.

Finally, the MIPS required must be calculated from the number of real operations required and the relevant parameters for each of the standards under consideration. Since multiplications (real or complex) are more demanding than additions, the complexity requirement is evaluated in two ways. In one case only the multiplications are considered – the additions being neglected. In the other case, both additions and multiplications are considered.

VI.1. REAL OPERATION REQUIRMENT

A complex addition can be implemented as two real additions. Generally, a complex multiplication can be implemented as three real multiplications plus five real additions [6]. In the case of the FFT, use of efficient radix-2, radix-4 or radix-8 algorithms can be used to reduce the number of non-trivial operations required [7]. Thus, the computational requirement to implement a PFE and a training algorithm is given by:

$$N_{RMULT-T} = 3(N_{CMULT-T} + N_{CMULT-F}) \qquad (12)$$

$$N_{OPS-T} = 2(N_{CADD-T} + N_{CADD-F}) + 8(N_{CMULT-T} + N_{CMULT-F}) \qquad (13)$$

The computational requirement to implement the PFE and a decision directed adaptation algorithm is given by:

$$N_{RMULT-DD} = 3(N_{CMULT-DD} + N_{CMULT-F}) + \frac{N_{RMULT-FFT}}{N} \qquad (14)$$

$$N_{OPS-DD} = 2(N_{CADD-DD} + N_{CADD-F}) + 8(N_{CMULT-DD} + N_{CMULT-F}) + \frac{N_{RMULT-FFT} + N_{RADD-FFT}}{N} \qquad (15)$$

VI.2. MIPS REQUIREMENT FOR LMS ADAPTATION OF THE PFE

Table 1 presents the relevant parameters of the three standards considered in this paper. Using these parameters in conjunction with equations 12, 13, 14 and 15, an initial estimate for the required number of MIPS for a given application can be determined. These estimates are summarized in table 2.

	HIPERLAN/2 / IEEE P802.11a	DVB-T 2k Mode
Tx Rate	20MHz	9MHz
Max. Delay Spread	800ns	50μs
OFDM Symbol Period	3.2μs	224μs
Guard Interval Fraction	1/4 or 1/8	1/4, 1/8, 1/16 or 1/32

Table 1: System Parameters

HIPERLAN/2: Figure 2 shows the computational requirements versus maximum delay spread capability for a HIPERLAN/2 or IEEE P802.11a PFE as a function of the number of equalizer taps, assuming LMS adaptation. A 9-tap PFE offers an efficiency increase of approximately 9% (since the guard interval can be reduced from 800ns to 400ns). This is equivalent to an increase of up to 4.8Mbits/s in raw data throughput. This is achieved at the cost of 1,680 additional MIPS (5,730 including additions) or 1,080 additional MIPS (3,600 including additions) if no decision directed channel tracking is employed. The latter case is a realistic option since the length of a HIPERLAN/2 burst is considerably shorter than the coherence time of the channel. Thus, the PFE offers a moderate increase in efficiency for a significant increase in complexity.

It should be noted that larger increases in efficiency could be achieved if the optional use of shorter guard intervals was supported by the standard.

IEEE 802.11a: Similar conclusions can be drawn for application of the PFE to IEEE 802.11a. However two important differences should be noted. Firstly, the optional short guard interval available in HIPERLAN/2 is not supported by IEEE 802.11a. The PFE will not actually offer an efficiency improvement for IEEE 802.11a unless an optional short guard interval is made available. Secondly, the potential for much longer transmission bursts exists in IEEE 802.11a. As a result, a PFE implemented in an IEEE 802.11a system is more likely to require decision directed tracking to accommodate channel variation over the burst duration.

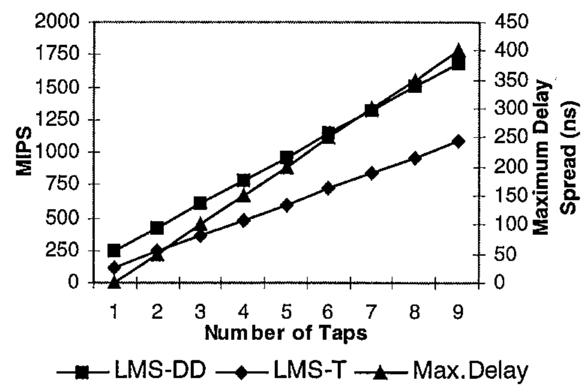

Figure 2. Complexity and Maximum Delay Spread of an LMS Adapted PFE for HIPERLAN/2 or IEEE 802.11a.

DVB-T 2k mode: Figure 3 shows the computational requirements versus maximum delay spread capability for a DVB-T 2k PFE as a function of the number of equalizer taps, assuming LMS adaptation. This system requires considerably more computational capability to achieve any useful PFE implementation. Although the lower transmission rate of DVB-T requires only 9/20 of the equalizer clock rate in comparison to HIPERLAN/2, the longer delay spreads in the outdoor broadcast channel results in the need for many more filter taps. Decision Directed channel tracking will almost certainly be required in this application. In this case, 20,800 MIPS (69,200 including additions) are required to achieve the same 9% efficiency improvement as for HIPERLAN/2. If only tracking is employed, 13,800 MIPS (46,100 including additions) are required. In either case, this represents a very poor complexity/efficiency tradeoff.

The application of the PFE to DVB-T 2k offers the potential to move from a multi-frequency network to a single frequency network. In this case, a potential seven-fold or even nine-fold increase (depending on the frequency re-use of the multi-frequency network) in spectral efficiency could be achieved. However, although this spectral efficiency improvement is extremely desirable, the complexity using the LMS algorithm requires in excess of 124,000 MIPS (415,000 including additions). This is clearly unacceptable.

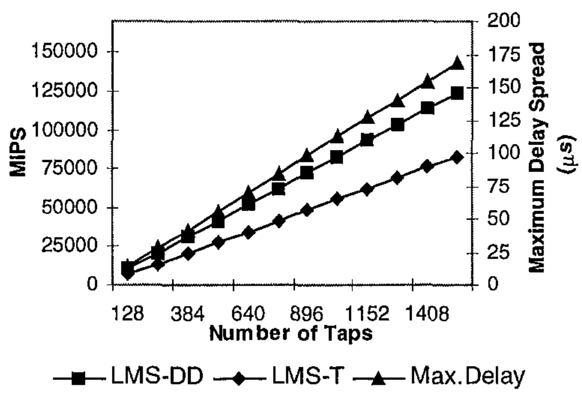

Figure 3. Complexity and Maximum Delay Spread of an LMS Adapted PFE for DVB-T 2k

Clearly, if an acceptable trade off between efficiency gain and complexity increase is to be achieved, alternative adaptation methods to the LMS algorithm must be considered.

	MIPS Requirement	
	Multiplies	Multiplies & Adds
HIPERLAN/2 or 802.11a Training	1,080	3,600
HIPERLAN/2 or 802.11a Decision Directed	1,680	5,730
DVB-T 2k Training	83,000	276,000
DVB-T 2k Decision Directed	124,000	415,000

Table 2: MIPS Requirements of an LMS adapted PFE.

VII. A CSI BASED COEFFICIENT CALCULATION METHOD

In section II, a combined OFDM-equalization receiver supporting an error based iterative adaptation algorithm was described. In this section, an alternative receiver structure (shown in figure 4) which supports CSI derived filter coefficient calculation and the CSI-based coefficient calculation algorithm itself are presented and discussed.

The feedforward section of the receiver supporting CSI-based equalizer adaptation functions in a similar manner to that of the receiver supporting error based adaptation. However, the feedback section functions differently. In this case, the feedback section takes the CSI information as its input and generates the equalizer tap coefficient vector in one of two ways, either in a 'single shot' manner or iteratively.

Single Shot Coefficient Calculation: The equalizer tap coefficient vector can be calculated from a single received OFDM symbol given the following conditions:

1. The received OFDM symbol is a training symbol (corresponding to a pilot symbol on all frequencies).
2. The receiver has a-priori knowledge of the transmitted OFDM training symbol.
3. The transmitted, training OFDM training symbol incorporates a guard interval of suitable length to ensure that no ISI is perceived at the receiver.
4. The equalizer is in an initial state in which it exhibits a frequently flat response (i.e. the equalizer tap coefficient is required to be a unit impulse response).

If the above conditions are met then the channel estimator will generate a CSI vector that is an accurate estimate of the frequency response of the channel. In this case, the CSI-based adaptation algorithm simply inputs the CSI vector, $S(k,l)$ to the IFFT process. The tap coefficient vector is thus calculated according to:

$$c(j,l+1)=\frac{1}{N}\sum_{k=0}^{N-1}S(k,l)e^{\left(\frac{i2\pi jk}{N}\right)} \quad (16)$$

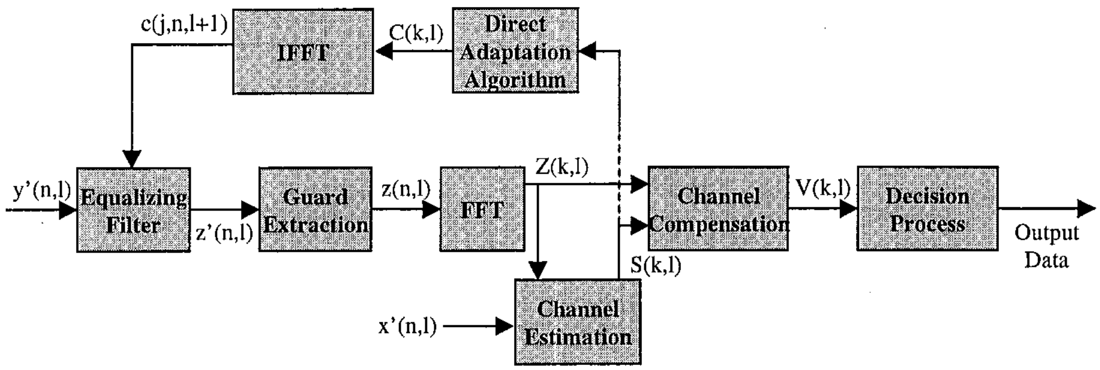

Figure 4. Combined OFDM-Equalization Receiver Supporting CSI-Based Adaptation

Thus an equalizer tap coefficient vector of length up to N taps can be generated. Using this approach, the equalizer tap coefficient vector can be calculated from a training sequence sent at the start of a transmission sequence.

Iterative Coefficient Calculation: The equalizer tap coefficient vector can also be updated in an iterative fashion. Iterative adaptation is possible provided that:

1. Adequate pilot symbols are inserted into the transmitted OFDM symbol to support the channel estimation process.
2. The received OFDM symbol has not suffered any ISI. This can be ensured either by means of a guard interval or if the PFE is already well adapted.
3. The receiver has a-priori knowledge of the inserted pilot symbols.

The criteria for iterative calculation of the equalizer tap coefficient vector are thus less stringent than those for single shot calculation.

If the above conditions are met then the channel estimator will generate a CSI vector that is an accurate estimate of the combined frequency response of the channel and the PFE. In this case, the CSI-based adaptation algorithm is required to multiply the current value of the CSI vector by the corresponding value of the CSI vector for the previous OFDM symbol period. The result is input to the IFFT process. The tap coefficient vector is thus calculated according to:

$$c(j,l+1)=\frac{1}{N}\sum_{k=0}^{N-1}\left(S(k,l),S(k,l-1)\right)e^{\left(\frac{i2\pi jk}{N}\right)} \qquad (17)$$

Using this approach, the equalizer tap coefficient vector can be calculated in an iterative manner from a series of OFDM symbols that contain a mixture of pilot information and data. The iterative CSI-based adaptation method is thus suitable both for the initial training of a PFE (provided that the a guard interval is employed until such time as the PFE is accurately adapted to the channel) or for 'fine tuning' of a PFE previously adapted by the single shot method (in which case no guard interval is required in the transmitted OFDM symbols).

VIII. COMPLEXITY ANALYSIS FOR THE CSI-BASED ADAPTATION METHOD

The complexity requirements for single shot and iterative CSI-based equalizer tap coefficient calculation can be considered separately. The total complexity required to implement a PFE and the appropriate adaptation strategy can then be determined.

Single Shot Complexity: The additional computational complexity of the single shot CSI-based adaptation method is simply that of the additional IFFT process. Thus, the total complexity requirement of the PFE and adaptation algorithm is given by:

$$N_{RMULT-CSI-S} = 3N_{CMULT-F} + \frac{N_{RMULT-FFT}}{N} \quad (18)$$

$$N_{OPS-CSI-S} = 2N_{CADD-F} + 8N_{CMULT-F} + \frac{N_{RMULT-FFT} + N_{RADD-FFT}}{N} \quad (19)$$

Iterative Complexity: The additional computational complexity of the iterative CSI-based adaptation method is the sum of the IFFT process and the N complex multiplications of the current and previous CSI vectors. Thus, the total complexity requirement is:

$$N_{RMULT-CSI-I} = 3N_{CMULT-F} + \frac{3N + N_{RMULT-FFT}}{N} \quad (20)$$

$$N_{OPS-CSI-I} = 2N_{CADD-F} + 8N_{CMULT-F} + \frac{10N + N_{RMULT-FFT} + N_{RADD-FFT}}{N} \quad (21)$$

IX. COMPLEXITY EVALUATION FOR THE CSI-BASED ADAPTATION METHOD

The complexity requirement for implementation of the PFE using CSI-based adaptation in each of HIPERLAN/2, IEEE 802.11a and DVB-T can now be evaluated. These requirements are summarized in table 3.

HIPERLAN/2: The training sequence preceding all bursts in HIPERLAN/2 is suitable for use with a single shot, CSI-based adaptation method. Figure 5 shows the computational requirements versus maximum delay spread capability for a HIPERLAN/2 or IEEE P802.11a PFE as a function of the number of equalizer taps, assuming CSI-based adaptation. The required computational complexity to implement single shot CSI-based adaptation of a 9-tap PFE in a HIPERLAN/2 system is 604 MIPS (2130 MIPS including additions). The computational complexity required to implement the iterative CSI-based adaptation method is 653 MIPS (2260 MIPS including additions). However, this is unlikely to be required since the length of transmission bursts in HIPERLAN/2 is typically less than the channel coherence time.

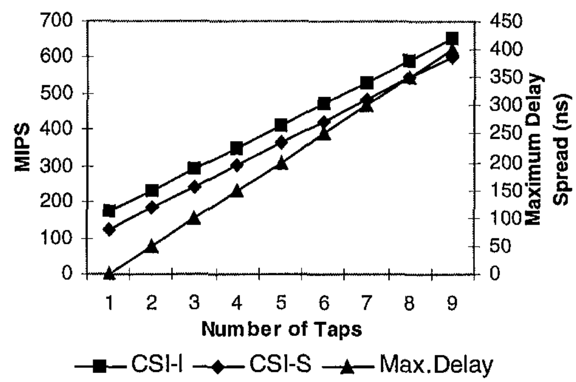

Figure 5. Complexity and Maximum Delay Spread of a CSI Adapted PFE for HIPERLAN/2 or IEEE 802.11a.

IEEE 802.11a: Similar conclusions to those made in section VI.2 are also applicable in this case.

DVB-T 2k: Figure 6 shows the computational requirements versus maximum delay spread capability for a DVB-T 2k PFE as a function of the number of equalizer taps, assuming CSI-based adaptation. Since DVB-T is a continuous broadcast system, iterative adaptation may be required. However, for this application, the difference between the requirement for single shot and iterative CSI adaptation is extremely small. In either case, the required computational complexity to allow DVB-T 2k operation in a single frequency network is approximately 41,500 MIPS (approximately 138,000 MIPS including adds).

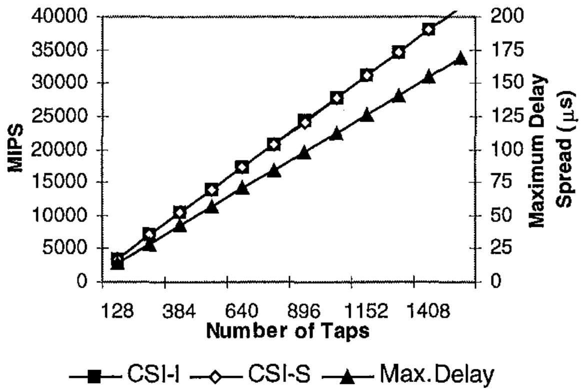

Figure 6. Complexity and Maximum Delay Spread of a CSI Adapted PFE for DVB-T 2k.

	MIPS Requirement	
	Multiplies	Multiplies & Adds
HIPERLAN/2 or 802.11a Single Shot	604	2,130
HIPERLAN/2 or 802.11a Iterative	653	2,260
DVB-T 2k Single Shot	41,500	138,000
DVB-T 2k Iterative	41,600	138,000

Table 3: MIPS Requirements of a CSI adapted PFE.

X. CONCLUSIONS

The computational complexity required to implement the PFE filter is high in the case of HIPERLAN/2 and IEEE 802.11a and extremely high is the case of DVB-T.

The evaluation of the complexity requirement for an LMS adapted PFE has shown that the complexity requirement for the implementation of the adaptation algorithm was a very signification part of the overall computation cost. Approximately 50% in the case of an equalizer adapted by training only and around 66% in the case of a system capable of decision directed equalizer adaptation. This is the case for LMS adaptation of a PFE for all three standards

The evaluation of the computational requirement of the CSI-based adaptation method indicated that this algorithm required a much lower fraction of the total computational cost. In the case of HIPERLAN/2 and IEEE 802.11a the CSI adaptation requirement represents around 11% of the total complexity for the single shot method and 17% of the total complexity for the iterative method. In the case of DVB-T the CSI adaptation requirement is a negligible fraction of the total. This is unsurprising, since the CSI-based method is specifically designed for OFDM systems, whereas the LMS algorithm and other error based iterative algorithms are generic.

Using the CSI-based adaptation method a low computational overhead for the adaptation of the PFE can be achieved. The computational cost of the PFE filter itself is not reduced. In order to achieve a more cost and complexity effective implementation of the PFE, methods to reduce the computational cost of the filter are required. This is less critical in HIPERLAN/2, where the required complexity may be more readily implemented. However, if the PFE is to be applied to DVB-T, the cost of the equalizing filter must be reduced considerably. Techniques being pioneered for application to the ATSC 8-VSB standard should be considered. Most 8-VSB products depend on an adaptive equalizer (typically a DFE) to combat ISI. These equalizers will typically implement a number of taps of the same order of magnitude as that required for DVB-T and will thus require similar computational complexity. Novel techniques pioneered for low cost implementation of equalizers for 8-VSB should be considered in terms of the suitability for application to combined OFDM-equalization for DVB-T.

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