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A Channel State Information Generation and Application Method In DVB-T System

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ABSTRACT

This paper presents the channel state information(CSI) generation and application method using the pilots inserted in transmitter in COFDM(Coded Orthogonal Frequency Division Multiplexing) system which is the standard transmission system of terrestrial digital TV in Europe. The pilots which are inserted in transmitter are used to synchronize the frequency and time and equalize the received carrier data in OFDM and using the CSI in receiver is a unique feature in OFDM compared to single carrier modulation. The CSI can be defined as a signal power(SP) or signal-to-noise ratio(SNR) in each carrier in COFDM. This CSI can in principle be applied for any selective channel and whatever modulation scheme is used on every carrier. In multi-carrier such as CODFM, the pilots are used as a reference signal in transmitter so these pilots can be used to derive the CSI. We have tested the CSI generation and application method of this paper under the standard Rayleigh(P1) and Ricean(F1) fading channel and analyzed the CSI generation method and application method to demapper. From the simulation, we obtained better performance than conventional soft decision and demapping method.

I. Introduction

In the European digital video broadcasting (DVB) project founded by initiative of more than 190 organizations, the standards for digital television broadcasting have been specified for different transmission media such as satellite (DVB-S[2]), cable (DVB-C[3]), terrestrial (DVB-T[1]). Compared with single carrier transmission system, COFDM modulation system employs multiple orthogonal subcarriers for the transmission of a parallel data stream. Because coded (usually concatenated code, that is, RS code and convolutional code, is used) and interleaved data are modulated on the multiple carriers, COFDM modulation system is more tolerant than single carrier transmission system under the multipath channel which can be viewed in the frequency domain as a frequency selective channel. Also COFDM modulation system can be used in so-called Single Frequency Networks (SFNs) in which all transmitters radiate the same signal on the same frequency. This SFN is a another feature of OFDM. For these reasons, COFDM modulation system is adopted as digital broadcasting systems in for both audio and video-DAB (Digital Audio Broadcasting[4]) and DVB-T (Digital Video Broadcasting-Terrestrial[1]) respectively.

Generally, when data are modulated onto a single carrier in a time-invariant system then a priori all data symbols suffer from the same noise power on average. So in case of single carrier modulation system the reliability of the received signal for soft decision is represented by only value proportional to distance from the decision boundary.

But, in the COFDM modulation system there is one more factor, known as CSI(Channel State Information)[5]. When data are modulated onto multiple carriers, as in COFDM system, the various carriers will have different SNR(Signal to Noise Ratio). For example, the carrier which falls

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in a frequency null is much noisier than one in a peak. The data which is modulated onto a carrier having high SNR is more reliable than the one having low SNR. This extra reliability information is called CSI and must take account of in the soft decision. Therefore, Viterbi decoder which has the input using the CSI is very important and unique configuration in the COFDM system.

To generate CSI, we have to estimate the SNR of each carrier position. Conventional SNR estimating method is to estimate signal power and noise power separately and then divide estimated signal power with noise power. But implementing this method is very complex and difficult. In this paper, we present a simple CSI generation and application method to demapper that uses CSI in receiver. This method does not require signal power estimation and noise power estimation to generate CSI. The organization of this paper is as follows. In Section II, OFDM system description and other related synchronization method are presented and in Section III, proposed CSI generation method is presented and in Section IV, CSI application method to demapper using CSI is presented. In section V, simulation results are provided to evaluate the performance. Finally, conclusions are in Section VI.

Ⅱ. OFDM System Description

In this paper, we considered two channel model that is Ricean fading (fixed reception : F1) and Rayleigh fading(portable reception : P1). A Rayleigh fading environment assumes no line-of-sight and no fixed reflectors/scatters. The expected value of this fading is zero. If there is a line-of-sight, then this fading can be modeled by Ricean fading. But this fading model has the same characteristics as the Rayleigh fading, except for a non-zero expected value. In this paper, we used the 20 multipath fading model[1] and the inter-carrier spacing of the system has to be chosen large compared to the maximum Doppler frequency of the fading channel. The complex data symbols xk are modulated on N subcarriers by an inverse discrete Fourier transform(IDFT) and the last L samples are copied and put as a guard interval to form the OFDM symbol. This data vector is serially transmitted over a discrete-time channel, whose impulse response is shorter than L samples. At the receiver, the guard interval is removed and the received signal rk is demodulated with a discrete Fourier transform (DFT).

Then, Lets consider this papers system. The insertion of a guard interval avoids ISI and preserves the orthogonality between the carriers, resulting in as follow.

$$y_k = h_k + n_{k,} k = 0, \ldots, N-1 \qquad (1)$$

where h_k is the complex valued channel attenuation at the kth subcarrier and nk is a sample from an additive complex white Gaussian noise process. In spite of the loss of transmission power and bandwidth associated with the guard interval, the simple channel equalization for the multi-channel structure (1) generally motivates the use of the guard interval. The uncertainty in the dispersion by the channel thus shows up as a multiplicative distortion for each subcarrier. The receiver block diagram of this paper is presented in Figure 1.

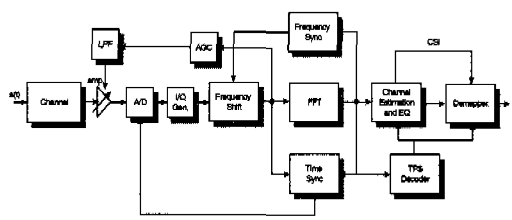

Fig. 1. OFDM Receiver Block Diagram.

1. *I-Q generation* : This part takes the input samples from the ADC output and convert this ADC output real-valued signal to complex-valued signal suitable for input to the FFT part by the Hilbert transform. This method based on the Hilbert transform performs better in situations of frequency instability or timing jitter. Above this reason, Hilbert transform method to convert the

real-valued signal to complex-valued signal is widely used. We also used this Hilbert transform to generate the complex-valued signal from the real-valued signal.

2. *Time synchronizationJ* : This part has the two purpose that one is to find the optimum timing for the starting position of the FFT window and the other is to synchronize the frequency of the ADC clock that has the 128/7 MHz to the received signal. The transmitter repeats a segment of the signal during the guard interval that is one of the among 1/4, 1/8, 1/16, 1/32 of the symbol duration. The scheme of finding the optimum timing for the FFT window is to use the guard interval inserted the transmitter part. Since OFDM symbols have the same data between the guard interval part and data part, the correlation is the generally used method to start position of the FFT.

3. *Frequency synchronization* : This frequency synchronization part is to correct for the tolerance in the frequency of the local oscillator in the tuner in digital domain and/or analogue domain. This part shifts the frequency of the complex baseband signal so that its center frequency is zero. The shift needs to be accurate to within a small fraction of the carrier spacing so that inter-carrier interference is kept to an acceptably low level and so that the tracking ability of the channel estimation is not absorbed in correcting for the error in the frequency of the local oscillator. This part comprises two parts. One is to shift the frequency of the complex baseband signal and the other is to measure the error in the center frequency of the complex baseband signal. The method of frequency synchronization is to shift the frequency of the complex baseband signal by multiplying the signal by a rotating vector. The error in the frequency of the local oscillator translates directly into an error in the center frequency of the complex baseband signal. The error may be several times the carrier spacing. So correction of the frequency error of local oscillator is divided into two part. One is the coarse part that the error is taken to the nearest multiple of the carrier spacing. The other is that the error of frequency is to be in range-0.5 to 0.5 of the carrier spacing.

The general method to measure the frequency error is to use the continual pilot carrier. The coarse frequency error is obtained by looking for the pattern of continual pilots. The effect of a fine error is to rotate the phase of all the received carriers by the same amount between two symbols. So fine error is obtained by measuring the phase rotation on continual pilots between one symbol and next symbol.

4. *Channel estimation and equalization* : The transmitted DVB-T signal contains scattered pilots which are distributed among the data carriers in a regular pattern. These scattered pilots are transmitted with known values, that is, imaginary part of pilots is always zero and real part of pilots has a fixed amplitude but with the sign determined from the carrier index number according to a PRBS function. So in this part, we used the this phenomenon. By comparing each received scattered pilots with the known transmitted value that is generated from the PRBS generator, a snopshot is obtained of the response of the channel for the corresponding carrier at that time instant. The data cells that must be corrected lie between the scattered pilots. So corrections are generated for each carrier using the interpolation of the scattered pilots. In this paper, we used the 23 tap filter to interpolate the pilots. The interpolator also slightly reduces the effects of thermal noise on the scattered-pilot measurement. After the estimation of channel response by the interpolation of frequency and time direction, the data carriers are divided by the estimated value from the interpolation. To do this way, channel estimation and equalization is easily done without any problem.

Ⅲ. CSI Generation

Frequently, CSI is defined as SNR in a carrier position. In the case of channel with only AWGN(Additive White Gaussian Noise), signal power (SP) estimation is sufficient enough to calculate CSI. But in the case of channel with frequency selective noise or narrow band interfering signal within the signal bandwidth, noise power differs from subcarrier to subcarrier and noise power estimation becomes essential for the performance improvement. However, the difficulty is that noise estimation isnt easy and transparent compared with signal power estimation that can be calculated in the channel estimation part of equalizer. Moreover, SNR calculation requires combining estimated signal and noise power in a constant rule. This means that first we have to estimate signal power from channel estimation part of equalizer and next we have to estimate noise power of each pilot carrier positions and then divide estimated signal power with estimated noise power. In designing hardware it is very difficult to design divider.

So, In this paper, we present a one-shot and simple SNR estimation method based on the MSE (Mean Square Error) estimation of the equalized output at the position of pilot carrier. Since we know the value of pilot carriers, we can calculate the error of the equalized output at the position of pilot carrier. We can assume that the channel is varying slowly enough to attain the reasonable MSE estimation in the time axis and the MSE at a pilot carrier position can be calculated by time-averaging the squared errors.

Assuming k as the position index of the carrier which its value is known (*i.e.* scattered pilot), we can estimate equalized output ek of the kth pilot carrier position as equation (2).

$$e_k = \frac{(p_k h_k + n_k)}{h'_k} \tag{2}$$

where p_k is the reference value of the pilot carrier, h_k is the channel response, h_k is the estimated value of the channel response h_k which can be obtained from channel estimator and n_k is noise component.

The MSE of e_k can be estimated as following equation.

$$MSE(e_k) = \langle\, |p_k - e_k|^2 \rangle = \langle\, |p_k(1 - \frac{h_k}{h'_k}) - \frac{n_k}{h'_k}|^2 \rangle \tag{3}$$

where < > denotes time-averaging function.

Equation (3) shows that when $h'_k \approx h_k$, MSE of e_k becomes the inverse of the signal to noise ratio. Therefore, we can utilize MSE estimation of e_k as SNR calculation. Moreover, in special case like DVB-T system, if the pilot signal p_k has only real value, with equation (2), equation (3) can be reduced to equation (4), provided that real and imaginary parts of n_k satisfy *i.i.d.* (independent and identically distributed) condition. And this shows that we dont need p_k any more.

$$MSE(e_k) = \langle\, (img[e_k])^2 \rangle \tag{4}$$

where $img[e_k]$ is the imaginary part of e_k.

Consequently, CSI can be generated as following manners. First, calculate MSE in the pilot positions by using equation (3) or (4). Secondly, calculate the inverse of MSE in that positions and normalize result value. Thirdly, interpolate above calculated values to get CSI in the useful data positions by using 0th or 1st order interpolation scheme.

Ⅳ. Demapping Method using CSI

In frequency selective channel, a carrier in a peak of the frequency response will be boosted and some which falls into a notch will be attenuated. In this case the CSI value from the estimator will also fluctuate. Because the CSI is the value of SNR, the equalized data having high CSI is more reliable than data having low CSI. In multipath channel like Ricean and Rayleigh fading channel, the variance of CSI is very high. So we have to limit the CSI value above a certain threshold, where is CSI value measured in the AWGN channel with single path. We assume data having CSI value above threshold as most reliable and make n-level reliability value by quantizing CSI value in the range 0 to in uniform step. Also the metric of the certainty of each decision m_i (for i th bit of a symbol of QAM signal) can be measured by the distance from the decision boundary of the signal constellation. Figure 2 shows 16-QAM constellation mappings and corresponding bit patterns of the DVB-T system[1].

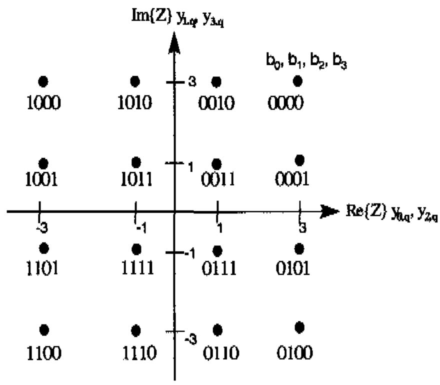

Fig. 2. 16-QAM Constellation Bits Pattern.

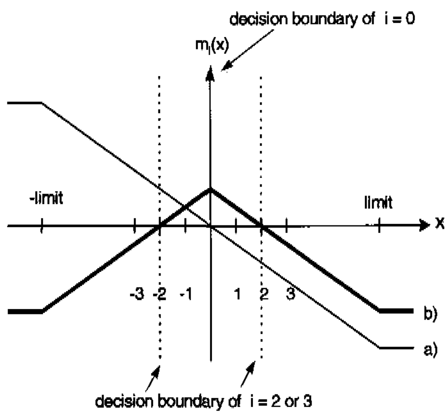

Fig. 3. Metrics of Each Bits.

Figure 3 shows metric of the certainty of the decision in demapper, in the case of the 16-QAM constellation of the DVB-T system. As you can see from Figure 3, graph a) is the metric value of the first (b_0) and third (b_2) bit in case of 16-QAM signal demapper and graph b) denotes the metric value of the second (b_1) and fourth bit (b_3), decision boundary of b_0 and b_2 of 16-QAM symbol are zero and the decision boundary of b_1 and b_3 are +2 and -2. Because of the property of QAM signal the priority of the b_0 and b_2 is higher than the priority of the bit $b1$ and $b3$. This property is used in generating the metric of the certainty value.

For example, lets assume that received 16-QAM symbol z has the value Real$\{z\}$=3.5 and Imag$\{z\}$=0.5. By using figure 3, the generated metric value of the first bit m_0 is larger than the metric value of the second bit m_1 . This results shows that b_0 is more reliable than b_1 in making decision that the value of the first bit is zero. So as you can see in Figure 3, instead of saturating the value $m_0(x)$ = b in region $|x|>1$ for the decision of the first bit and saturating $m_3(x)$ = b in region $|x|<1$ and $|x|>3$ for the decision of the third bit, it is more reasonable to extend the $mi(x)$ value linearly to the input value x.

Because both CSI and m_i are directly proportional to the reliability, we can obtain the final reliability value r of the decision by just multiplying CSI by m_i as equation (5).

$$r_i = CSI_k * m_i \tag{5}$$

where $CSIk$ is CSI value of k th carrier position and i is the bit index of the carrier. For Viterbi decoder the reliability value r must be quantized appropriately.

V. Simulation Results

The simulations are performed under the Mode 2 (16-QAM, 1/2 code)[10]. DVB-T standard[1] defines many different transmission modes by selecting the inner code rate (1/2, 2/3, 3/4, 5/6 and 7/8 inner code rate), different modulation such as QPSK, 16-QAM and 64-QAM, guard interval such as 1/4, 1/8, 1/16 and 1/32 of the symbol duration and the number of carrier that is 1705 carriers or 6817 carriers. For some of the combinations the C/N (carrier-to-noise) ratios required to obtained a BER (Bit Error Rate) = 2 $*10^{-4}$ are provided in the standard specifications as indicative values. This BER means that Quasi Error Free (QEF) performance after the Reed-Solomon decoding. The simulations are performed with the DVB-T system which has the 1705 subcarriers, 1/2 code rate, non-hierarchy and 16-QAM constellation with guard interval 1/32. The Ricean (F1) and Rayleigh (P1) channel parameters are taken from DVB-T specification[1]. In the simulation, we used the Viterbi decoder that has 96 truncation depth. The BER is measured after 4 bit soft decision Viterbi decoder. We calculated the performance difference at BER $2*10^{-4}$.

We made a simulation of 3 super-frame input data, that is, in 16-QAM in DVB-T, 1 super-frame input is 504 RS encoded Transport Stream (TS) packet so the total number of transmitted data of this simulation is 504*204*8*3 = 467584 bits. And we used Viterbi decoder that has the 96 truncation depth. This 96 truncation depth is widely used truncation depth in most commercial Viterbi decoder ICs so we used 96 truncation depth in Viterbi decoder. Also Viterbi decoder has a 4 bits soft decision input signal which has the signed magnitude value.

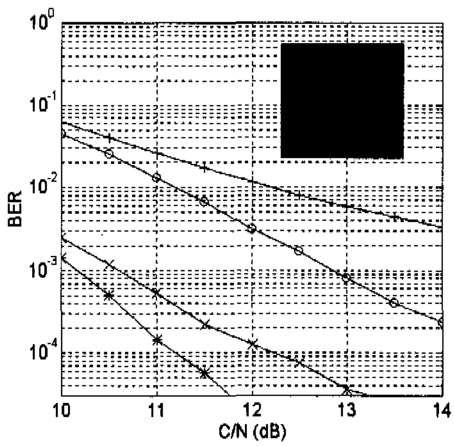

Fig. 4. BER Comparision of SP and SNR.

Figure 4 shows the simulation results in the frequency selective noise channel where C/I (Carrier to Interference noise ratio) is 15 dB and bandwidth of the inserted Gaussian noise interference is about 0.9MHz. Here we compared a demapper using only signal power (SP) as CSI with a demapper using SNR as CSI. In Ricean channel (F1) we obtained about 0.8 dB performance improvement. In Rayleigh channel (P1), if SNR is not considered as CSI then the BER of the system can not reach $2*10^{-4}$.

From figure 4, we can see that CSI as a SNR is much superior to CSI as a only signal power (SP).Figure 5 shows the performance of the demapper between *mi*(*x*) saturated method and *mi*(*x*) extended method. In Ricean (F1) channel, we obtained the about 0.35dB better performance and in Rayleigh (P1) channel, we obtained the about 1.1dB better performance by using the extended method. So we can see that extended method is much superior to saturated one.

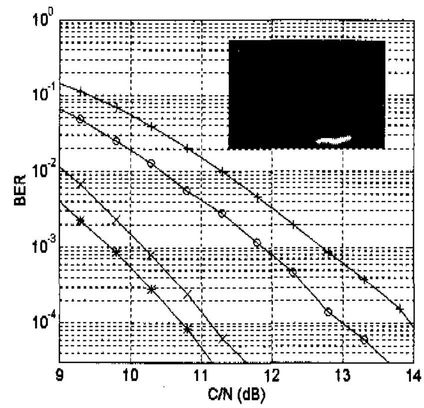

Fig. 5. BER Comparision of Saturated and Extended Method.

Ⅵ. Conclusion

In this paper, we have presented a simple and very efficient method of CSI generation and application method to demapper using CSI. Without estimating the noise power, SNR could be directly obtained by time averaging the MSE of the estimated pilot carriers. This CSI generation method is very efficient when there is any interference signal under fading channel or other channel and we could see the performance difference between CSI as a signal power (SP) and CSI as a signal-to-noise ratio (SNR).

Also this paper's CSI application method to demapper is superior to saturated scheme. Finally, by using the proposed CSI generation method and demapping method, we could confirm that the performance of the OFDM system, that is BER after Viterbi decoder, has been greatly improved in simulation under Rayleigh (P1) channel and Ricean (F1) channel.

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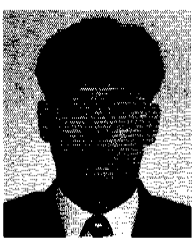

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