How to combat long echoes in OFDM transmission schemes: Sub-channel equalization or more powerful channel coding

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Abstract

The OFDM technique with guard-time, which has been proposed for terrestrial digital image and digital audio broadcasting applications in Europe, is an interesting approach to combat the frequency selectivity of the channel. However, it is sensitive to the interference caused by echoes longer than the guard-time. The aim of this article is twofold. First, the effects of the guard interval on a real case multi-path channel are analyzed for an OFDM transmission system. The system performance is analytically derived as a function of the guard interval. The sensitivity of the OFDM system to the interference caused by echoes exceeding the guard-time is analyzed. The degradation due to the interference can be restored by applying two different techniques: using more powerful channel coding or employing sub-channel equalization. As channel coding, a combination of multilevel coding with multiresoltion modulation, which is proposed for a digital terrestrial TV broadcasting [3], is used. The sub-channel equalizer is based on DFE, where its adaptation calls-upon the LMS-algorithm, using the reference symbols sent periodically. Simulation results show that using powerful channel coding, rather than equalization, can partly restore the degradation caused by a shorter guard-time.

1 Introduction

Orthogonal frequency division multiplexing (OFDM) is an efficient data transmission system, which is particularly suited for transmission over frequency selective fading channels. On such channels, the insertion of a guard interval before each transmitted block, longer than the largest delay spread, removes the necessity of further inter channel interference (ICI) equalization [1, 2].

A loss of $10\% - 25\%$ in spectral efficiency due to the insertion of the guard interval is commonly accepted [3]. For some multi-path channels having larger delay spread (e.g. single frequency network applications), this requires longer OFDM symbol duration with a corresponding increase in number of sub-carriers. This also entails higher FFT complexity and greater phase noise sensitivity [4]. The use of a fixed guard interval does not protect the system if the channel has longer echoes.

In general, equalization can partly compensate the losses in these cases and improves the system robustness. A time-equalization, using tapped delay line would require a large number of taps which results in slow convergence of the adaptation algorithm. Frequency-equalization was first presented by Weinstein and Ebert in '71 [5]. The idea was to use the least mean square (LMS) algorithm to independently adapt the one-tap equalizers at the output of the FFT (frequency domain). An immediate advantage of this structure is that no channel estimator is required. This first frequency-equalizer could only equalize distortion due to co-channel interference (CCI). To equalize ICI also, a more complex structure is needed. In [6] three tap equalizers per sub-channel are used to completely equalize the ICI of an orthogonally multiplexed staggered QAM.

The aim of this article is to study the dependence of the system performance from the guard interval duration and the benefits of coding and equalization. However, with OFDM transmission, complete ICI cancellation will require an N-tap equalizer for each sub-channel, N being the total number of sub-carriers. This equalization scheme is clearly impractical even for small values of N. To keep complexity low, only partial equalization will be considered. As far as channel coding concerned, we will study the performance of a powerful concatenated coding scheme, that has been proposed for terrestrial digital TV broadcasting in Europe [7]. It is based on a combination of multilevel coding and 64 Multi-resolution QAM (64-MRQAM), allowing great flexibility in the choice of the system parameters.

The article is organized as follows. The transmission system under study will be described briefly in section 2. In section 3 we analytically derive the system performance for a multi-path channel as a function of the guard interval duration. In section 4 the channel coding scheme and a simple equalization technique are presented. In section 5 simulation results are given. Section 6 is devoted to some conclusions.

2 An OFDM transmission scheme

Figure-1 illustrates the transmission scheme under study. It is proposed for terrestrial digital-TV broadcasting [3]. It consists of a channel encoder, modulator, OFDM multiplexer, D/A converter and a transmitter front-end. The receiver comprises a receiver front-end, an A/D converter, a synchronization mechanism, an OFDM demultiplexer, channel estimator, demodulator, and channel decoder. The source codec is based on MPEG-2.

The router classifies the different priority streams (HP, MP and LP) issued from the hierarchical source coding. The HP stream destinated for portable SDTV-reception is highly protected. This is also used as the basis for graceful degradation for stationary

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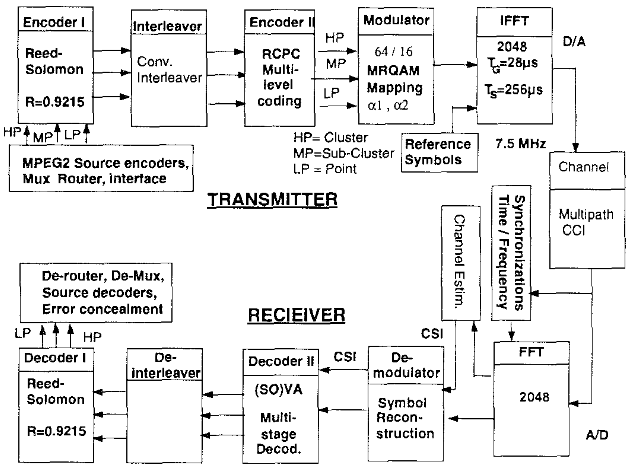

Figure 1: A Transmission scheme based on OFDM

EDTV or HDTV receivers.

The channel coding is based on a concatenated coding approach, where, the inner code is combined with Multi-resolution 16/64-QAM [7]. The outer code is a Reed Solomon code.

The OFDM technique with guard-time [2] is used to combat long echoes and the PAL/SECAM co-channel interference (CCI). The OFDM multiplexing and de-multiplexing will be done by a 2K point IFFT/FFT. The sub-carriers are spaced by $4883Hz$ (symbol-duration: 205 μsec). The number of useful carriers is 1536 (useful bandwidth 7.5 MHz).

The synchronization and the channel estimation are based on the known transmitted symbols and cells [3]. These are inserted in the OFDM frame in such a way that they can be used for different purposes: channel- and phase-noise estimation and fine time- and frequency-synchronization.

From the received signal after OFDM de-multiplexing and channel estimation, the corresponding demodulation and inner multi-stage decoding is performed depending on the receiver type [7].

The OFDM technique is particularly suited for transmission over frequency selective fading channels. The insertion of a guard-time T_g before each transmitted block, longer than the largest delay spread, removes the necessity of further inter channel interference (ICI) equalization [1, 2]. For some multi-path channels having larger delay spread (e.g. single frequency network applications), this requires longer OFDM symbol duration with a corresponding increase in number of sub-carriers. This also entails higher FFT complexity and greater phase noise sensitivity [4]. However, the use of a fixed guard interval does not protect the system if the channel has longer echoes. Let's look at these effects in more details.

3 Effects of guard-time in OFDM

Let $f_k = f_c + k/T_b$ for $k = 0 \ldots N-1$ be the N sub-carrier frequencies, where f_c is the carrier frequency and T_b is the effective OFDM *block* duration. Let T_g be the duration of the *guard interval* and $T_b' = T_b + T_g$ the total block duration. The orthogonal *basis functions* are:

$$\phi_k(t) = \begin{cases} e^{j2\pi f_k t} & -T_g \le t < T_b \\ 0 & \text{otherwise} \end{cases} \qquad (1)$$

and the time domain OFDM transmitted signal can be written as [2]

$$x(t) = \sum_{n=-\infty}^{+\infty} \sum_{k=0}^{N-1} X_{n,k} \phi_k(t - nT_b') , \qquad (2)$$

where $X_{n,k}$ are the QAM *symbols* transmitted on the k-th sub-carrier in the n-th block.

The time domain signal (2) is constructed in the modulator by using an N-samples IFFT followed by a D/A conversion (see Figure 1). The optimal decoding rule for the AWGN distortionless channel performs the integrate and dump operation [2] over the effective block duration T_b for each sub-channel $k = 0 \ldots N-1$, on the n-th block

$$X_{n,k} = \frac{1}{T_b} \int_0^{T_b} x(t) \phi_k^*(t - nT_b') dt = \frac{1}{T_b} \int_0^{T_b} x(t) e^{-j2\pi f_k (t - nT_b')} dt \qquad (3)$$

This operation is equivalent to a bank of matched filters followed by samplers, but it can be more efficiently implemented performing an FFT operation on the samples of the received signal [2].

3.1 OFDM in a Multi-path channel

We consider the transmission of the signal over a multi-path channel with impulse response [10]

$$h(t) = \frac{1}{\sqrt{N_p}} \sum_{m=1}^{N_p} e^{j(\theta_m + 2\pi f_{D_m} t)} \delta(t - \tau_m) , \qquad (4)$$

where N_p is the total number of reflected paths, τ_m, θ_m and f_{D_m} are delay, phase rotation and Doppler frequency of each path, respectively. These parameters can be derived from the channel *scattering functions* [10]. For large values of N_p this corresponds to the frequency/time selective Rayleigh fading channel. In the following we will assume that the Doppler frequencies are small compared to $1/T_b'$ (or $f_D T_b' \ll 1$), which is usually the case for portable or fixed receivers. Therefore, the phase rotation due to the Doppler-frequency can be replaced by a constant phase shift in the integration interval of each block. We note that the condition that $f_D T_b' \ll 1$ is the major limitation to the increase of the OFDM block duration T_b': Low immunity to phase noise due to the domestic local oscillator [4] and high sensitivity to the channel variations.

In the absence of noise the received signal before FFT is

$$y(t) = \frac{1}{\sqrt{N_p}} \sum_{m=1}^{N_p} e^{j(\theta_m + 2\pi f_{D_m} t)} x(t - \tau_m) , \qquad (5)$$

Let us now study, with the aid of a simple case, how the received signal is demodulated for different guard interval lengths. Linearity enables to separately consider the effects of each path. We can analyze first a simple case with a direct path and one echo (two path Rician fading)

$$y(t) = x(t) + A e^{j(\theta + 2\pi f_D t)} x(t - \tau) , \qquad (6)$$

where τ is the delay, θ the phase rotation, f_D the Doppler frequency and A the attenuation.

Then we can extend the results to the general channel model (4). The output of the k-th sub-channel for the n-th block can be written as the sum of two terms

$$Y_{n,k} = X_{n,k} + E_{n,k} \ , \tag{7}$$

the first term is due to the direct path, while the second one is the interfering term produced by the echoes of the preceding blocks

$$E_{n,k} = \frac{1}{T_b}\int_0^{T_b} Ae^{j(\theta+2\pi f_D t)}x(t-\tau)\phi_k^*(t-nT_b')dt \ . \tag{8}$$

If we assume that $\tau \leq T_b'$, then, only the $(n-1)$-th block will interfere with the n-th block and we only need to consider two cases: $\tau \leq T_g$ and $T_g < \tau \leq T_b'$.

3.2 Delay spread smaller than the T_g

If the delay spread $\tau \leq T_g$, then no echo from the $(n-1)$-th block will reach the integration interval for the n-th block and

$$E_{n,k} = \frac{Ae^{j(\theta+2\pi f_D nT_b')}}{T_b}\sum_{l=0}^{N-1} X_{n,l}e^{-j2\pi f_l \tau}\int_0^{T_b} e^{j2\pi(f_l-f_k)u}du \ . \tag{9}$$

Since

$$\frac{1}{T_b}\int_0^{T_b} e^{j2\pi(f_l-f_k)u}du = \begin{cases} 1 & l=k \\ 0 & l\neq k \end{cases} \tag{10}$$

we say that the *orthogonality conditions* are satisfied and

$$E_{n,k} = Ae^{j(\theta+2\pi f_D nT_b')}e^{-j2\pi f_k \tau}X_{n,k} \ . \tag{11}$$

Then, the received symbols are

$$Y_{n,k} = \left[1 + Ae^{j(\theta+2\pi f_D nT_b')}e^{-j2\pi f_k \tau}\right] X_{n,k} = H_{n,k}X_{n,k} \ , \tag{12}$$

where $H_{n,k}$ are the samples taken at the sub-carrier frequencies of the channel transfer function $H(f, nT_b')$ for the n-th integration interval. For the general channel model (4) we have

$$Y_{n,k} = \left[\frac{1}{\sqrt{N_p}}\sum_{m=1}^{N_p} e^{j(\theta_m+2\pi f_{D_m} nT_b')}e^{-j2\pi f_k \tau_m}\right] X_{n,k} = H_{n,k}X_{n,k} \tag{13}$$

As this expression shows, each sub-channel is only affected by the distortion within the same sub-channel, which we call co-channel interference (CCI). In this case, the transmitted information is completely recovered by multiplying the received symbols by the channel coefficients $H_{n,k}^{-1}$, if the channel is perfectly known. In practice the $H_{n,k}$ is estimated by sending periodically known pilot cells and performing an interpolation filter [3, 4]. The insertion of the pilot-cells should be performed in such a way that they can track slow channel variations produced by the Doppler frequencies.

3.3 Delay spread greater than the T_g

In this case, the orthogonality conditions do not hold any more. The interfering term consists of two parts: the first is produced by the echo of the preceding block and the second is due to the loss of orthogonality in the remaining part of the integration interval

$$E_{n,k} = Ae^{j(\theta+2\pi f_D nT_b')}\left\{\sum_{l=0}^{N-1} X_{n-1,l}\lambda_{l,k}(\tau) + \sum_{l=0}^{N-1} X_{n,l}\mu_{l,k}(\tau)\right\} \tag{14}$$

where, with some simple calculations,

$$\lambda_{l,k}(\tau) = \begin{cases} (\frac{\tau-T_g}{T_b})e^{j2\pi k(T_b'-\tau)/T_b}, & \text{for } l=k \\ e^{j\pi[2l(T_b'-\tau)/T_b+(l-k)(\tau-T_g)/T_b]}\dfrac{\sin[\pi(l-k)(\tau-T_g)/T_b]}{\pi(l-k)} & \end{cases}$$

$$\mu_{l,k}(\tau) = \begin{cases} (\frac{T_b-\tau+T_g}{T_b})e^{-2\pi jk\tau/T_b}, & \text{for } l=k \\ -e^{j\pi[-2l\tau/T_b+(l-k)(\tau-T_g)/T_b]}\dfrac{\sin[\pi(l-k)(\tau-T_g)/T_b]}{\pi(l-k)} & \end{cases}$$

We note that these coefficients depend only on τ and T_g.

By generalizing to the multi-path channel model given in (4), we obtain the received signal

$$Y_{n,k} = \frac{\mu_{k,k}}{\sqrt{N_p}}X_{n,k} + \sum_{l=0,l\neq k}^{N-1} X_{n,l}\frac{\mu_{l,k}}{\sqrt{N_p}} + \sum_{l=0}^{N-1} X_{n-1,l}\frac{\lambda_{l,k}}{\sqrt{N_p}}, \quad \text{for} \quad k=0\ldots N-1. \tag{15}$$

where

$$\lambda_{l,k} = \sum_{\tau\in\{\tau_m>T_g\}} B_m\,\lambda_{l,k}(\tau), \quad \text{for} \quad l\neq k \tag{16}$$

$$\mu_{l,k} = \sum_{\tau\in\{\tau_m>T_g\}} B_m\,\mu_{l,k}(\tau), \quad \text{for} \quad l\neq k \tag{17}$$

$$\mu_{k,k} = \sum_{\tau\in\{\tau_m\leq T_g\}} B_m\,e^{j2\pi f_k\tau} + \sum_{\tau\in\{\tau_m>T_g\}} B_m\,\mu_{k,k}(\tau) \tag{18}$$

and $B_m = e^{j(\theta_m+2\pi f_{D_m} nT_b')}$. To simplify the notation, the dependence from n has been omitted.

The information symbols can no longer be so easily extracted from $Y_{n,k}$ because of the interference of one block with the following. This causes inter channel interference (ICI) since some of the power transmitted in one sub-channel is transferred into the adjacent sub-channels during transmission. We note that this interfering power decreases with $sin^2(\pi(l-k)(\tau T_g)/T_b)/(l-k)^2$ as we move away from the main sub-channel. If we assume that the symbols of the preceding block $X_{n-1,l}$ for $l=0\ldots N-1$ are known (they have been correctly decoded in the previous interval), then (15) is a set of linear equations in the unknowns $X_{n,k}$, for $k=0\ldots N-1$. The solution of such a set of equations gives the optimum demodulator for the given channel. It is clear that the complexity of this demodulator becomes prohibitive even for small values of N.

A final remark on the type of interference is useful. In this case we could actually speak about inter-block interference which can not be absorbed by the guard interval. This interference results, in the frequency domain terminology, in both ICI and CCI.

Therefore, if we perform only the usual channel estimation procedure we find a systematic error term to the channel coefficient, due to the interference. The power of this error-term depends strongly on τ.

With the assumption that the transmitted symbols are statistically independent we then obtain the ratio of the signal power to the interference power at the output of each sub-channel as

$$(C/I)_k = \frac{|\mu_{k,k}|^2}{\sum_{l=0,l\neq k}^{N-1}|\mu_{l,k}|^2 + \sum_{l=0}^{N-1}|\lambda_{l,k}|^2}, \qquad k=0\ldots N-1 \tag{19}$$

4 Improving the system

If the guard interval is not sufficient to combat the channel selectivity, we must call upon to other techniques to achieve the desired performance: more powerful channel coding, sub-channel equalization or a combination of both.

In order to accelerate the convergence of the equalizer, we have to devote more capacity for reference symbols, and using a given channel coding. On the other hand, we can spend this extra redundancy to improve the performance of channel coding.

4.1 Adaptive equalization

The optimum demodulator is clearly impractical, so we must resort to some sub-optimum device which at least reduces the error produced by the echoes exceeding the guard interval duration. If we consider the interfering term in eq. (15), we note that the most relevant terms $(\lambda_{l,k}, \mu_{l,k})$ appear for small values of $l-k$. For $l \neq k$ the λ's and the μ's give the same contribute to the error term. While for $l = k$ the error depends on the actual delay spread of the channel. All these terms, except for $\mu_{k,k}$ account for the ICI in the sub-channel k produced by the nearest neighboring sub-channels.

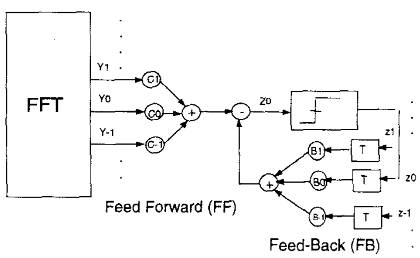

Figure 2: A Decision Feedback equalizer structure

This fact suggests the decision feedback equalizer structure shown in Fig.2, which has to be added to each sub-channel output after the FFT of Figure 1.

The $2 \times 2N_t + 1$ tap coefficients $C_{p,k}$ and $B_{p,k}$, $|p| \leq N_t$, $k = 0 \dots N-1$ have to be computed so that the mean square error at the output of each equalizer is minimum, that is

$$\min_{(C_{-N_t,k},\dots,C_{0,k},\dots,C_{N_t,k}),(B_{-N_t,k},\dots,B_{0,k},\dots,B_{N_t,k})} E\left\{|X_k - Z_k|^2\right\}, \quad (20)$$

where

$$Z_k = \sum_{p=-N_t}^{p=N_t} C_{p,k} Y_{n,k-p} - B_{p,k} Y_{n-1,k-p} \quad (21)$$

is the output of the k-th equalizer. The simplest adaptation algorithm converging to the optimal solution of (20) used here, is the LMS algorithm, also known as the stochastic gradient algorithm. In oder to accelerate the convergence of the equalization process, we have to send more reference symbols periodically. The extra amount of redundancy will be about 5-10 %.

Since the multi-path channel is frequency selective, the adaptation coefficients α_k will be different in each sub-channel. In the faded sub-channels the SNR can be very low and in such cases the adaptation algorithm may not converge. The complexity of the equalization process in terms of number of multiplications per received block is $2(2N_t+1)N + 2(2N_t+1)2N$. The first term accounts for the equalizer output calculation, while the second term accounts for the adaptation.

4.2 Powerful Concatenated channel coding

Another method to improve the system performance is to use more powerful channel coding, rather than equalization, by exploiting the extra redundancy spent for equalization.

We consider a concatenated channel coding scheme, where, the outer code is based on Reed-Solomon code. In order to achieve high spectral efficiencies, the inner multilevel coding and the Multi-resolution modulation are combined by using the principle of Set-Partitioning [7].

Let's denote C_i, with Hamming distance d_{H_i}, with code rate r_i, the code utilized for the i^{th} bit. The issued coded bits b_i, $i = 1, ..., m$, are then mapped to a 2^m-point MRM constellation. For our example of 64-MRQAM, the two first HP bits (b_1, b_2) are mapped to the clusters, then the two other MP bits (b_3, b_4) to the sub-clusters and finally the last two LP bits (b_5, b_6) choose a point of these sub-clusters [7]. This results in a hierarchical coding scheme.

This combination allows one to obtain high spectral efficiencies and to adapt the codes and the modulation to different reception conditions by changing the constellation ratios (α_1 and α_2) and the Hamming distances d_{H_i} of the different codes.

5 Simulation results

The channel model that we have considered is a conventional network in hilly terrain. It is a frequency selective fading channel model, where, its delay spread is 87 μsec with three main groups having 33%, 50% and 17% of the total energy, respectively. In Figure 3 the channel transfer function without any interference is plotted, where, some sub-channel may have less than 25 dB power than the average power of 0. dB.

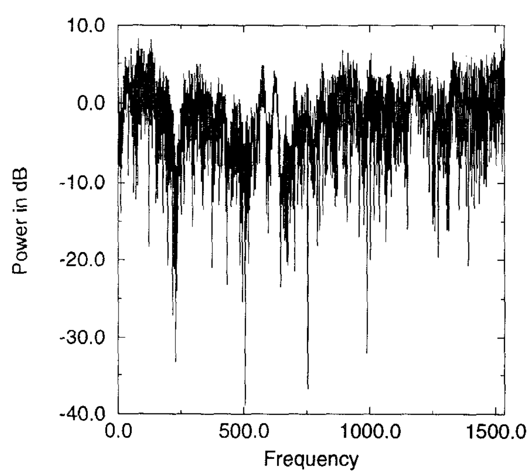

Figure 3: Conventional network channel model for H.T.

We consider two disturbances: Rayleigh fading channel as a portable receiver (SDTV) condition and Rician fading as a stationary receiver (EDTV, HDTV) condition.

5.1 C/I and the power of interferences

For the above channel by simulations we draw the average C/I (averaged over all sub-channels without noise) as a function of the guard-time T_g which is given in Figure 4. One can see that for large guard time the C/I is quite high: more than 30 dB. However, for no-guard time case, the C/I goes down to less than 10 dB, since the echoes can enter 40 % of the symbol duration.

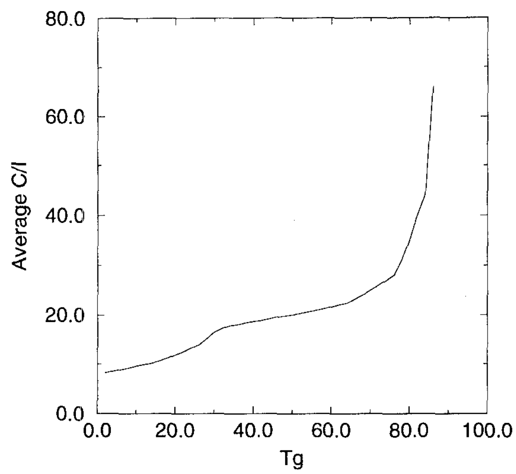

Figure 4: The average C/I vs. T_g

The interference-terms for this channel (assuming that the delay, phase and attenuation for each path are known) for each sub-carrier can be analytically derived, using the expressions given in section 3. For each sub-channel the interference term is made out of two part: The first part is coming from the same block (because of the loss of the orthogonality) and the other term comes from the preceding blocks. Here, the delay spread is lower than T_b', hence, this last term comes only from the last block.

In Figure 5, we have plotted the amount of interference-power coming from the other sub-carriers to the 50th sub-carrier considered as a reference (this reference frequency has a power of -1.5 dB), as a function of sub-carrier number (frequency), for $T_g = 40\mu\, sec$. In this Figure both terms of interferences (the interference coming from the same (ISB) and the preceding blocks (IPB)) are plotted. One can see that first of all some symmetry around the 50th sub-carrier exists and in addition ISB and IPB have relatively the same contribution to the 50th sub-carrier. However, the power of their contributions decreases as we are far from the reference carrier.

5.2 DFE and the optimal number of taps

The interefernce contributions imply a decision-feedback-equalization (DFE) structure where, the total number of coefficients N_c are equally distributed for the feed-forward (FF) and for the feed-back (FB) loops. Then, the required number of taps to cut the interference power to a certain level, for instance less than 20-25 dB can be given by above interference terms. However, the optimal case (with no residual interference) would require at least 1536 coefficients, which is not-realizable.

In Figure- 6 the required number of taps for guard-time $T_g =$ 70 μsec is given. As this figure show, even for this large guard-time, one need a high number of taps.

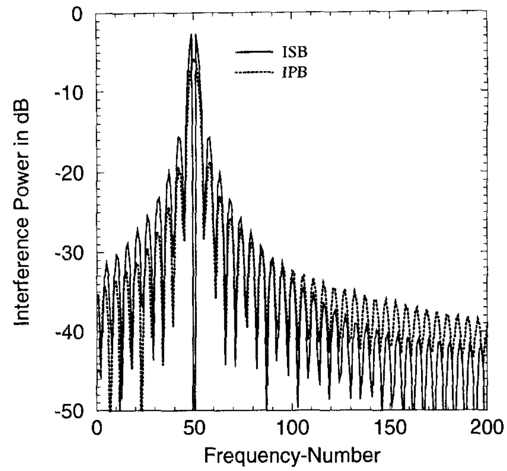

Figure 5: The interference contributions $T_g = 40\mu$sec

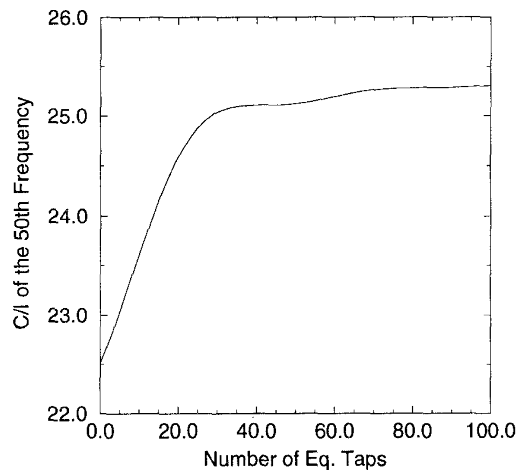

Figure 6: Number of Taps for $T_g = 70\mu$sec (50th sub-carrier)

A DFE structure have been simulated, where the adaptation is based on LMS algorithm. 10 % of the capacity was devoted for reference symbols, sent periodically. The number of taps considered are for FF and FB loops independently are 3, 5. The speed of convergence is given by the parameter α which depends on each sub-carrier, since each sub-carrier has different power.

In the same time we have also simulated the case that we perform a simple channel estimation (CE) as usual, by averaging the coefficients estimated from the reference blocks. Then, the coefficients obtained can be used directly to the Viterbi decoder.

In Figure 7, the performance of the equalizer with inner channel encoder, for the LP-bit stream in a Rician fading channel (K=10), for different number of taps is given. The guard time $T_g = 60\mu sec$. The LP-bit stream is protected with code rate (r5=5/6,r6=7/8) derived from the mother code of rate 1/3 with 64 stated convolutional code. The same simulations for the other layers is per-

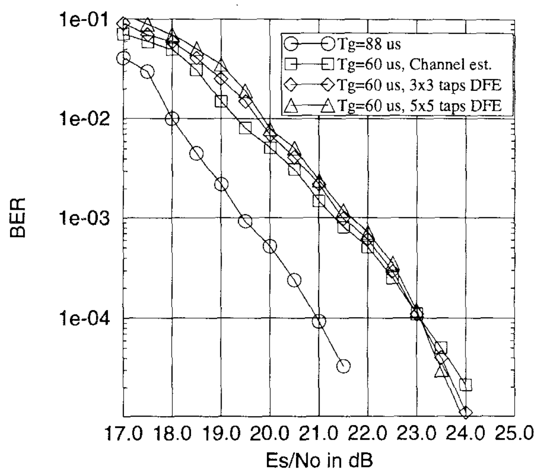

Figure 7: Performance of equalization and coding

	Table 1: Performance of channel coding vs. T_g				
P	Code rates	$T_g = 88$ inμsec	$T_g = 75$ in μsec	$T_g = 60$ in μsec	$T_g = 40$ in μsec
H	$r_1 = \frac{1}{3}, r_2 = \frac{2}{3}$	11.5	12.3	EF	EF
	$r_1 = \frac{1}{3}, r_2 = \frac{4}{7}$	10.8	11.5	13.5	EF
M	$r_3 = \frac{5}{6}, r_4 = \frac{7}{8}$	17.5	18.2	19.5	EF
	$r_3 = \frac{4}{7}, r_4 = \frac{4}{5}$	16.5	16.9	18.1	22
L	$r_5 = \frac{5}{6}, r_6 = \frac{7}{8}$	20.5	21.2	22.5	EF
	$r_5 = \frac{4}{7}, r_6 = \frac{4}{5}$	19.5	20.1	21.3	25

formed. From our simulations we can make the following remarks.

- The system may not converge in a Rayleigh fading channel.
- If the system converges, the convergence speed is very slow. In some cases, it needs more than 100- FFT blocks. The convergence speed depends on the total number of taps.
- The use of channel estimation for different T_g outperforms or has the same performance as the system with equalisation. Hence, there is no improvement by equalisation.
- The convergence may speed-up if one uses the channel-estimation values as the initial values of the central coefficients of the equaliser.

The interpretation of these results are the following. First of all since in some sub-channels we have very low power, the equaliser will try to equalise the noise, which may results in a noise amplification. Second remark, if the C/I is large enough, then the CE directly works better. Otherwise for low C/I, we need a huge number of taps, where its good-convergence time will be much longer than 100-200 blocks, which is not a realistic case.

5.3 More powerful channel coding

The 10% of the redundancy spent for equalization, will be used for channel coding. As we use powerful Reed-Solomon outer code, a bit-error-rate (BER) of 4.10^{-4} is sufficient to cut down to a BER less than 10^{-10}, that is required for image transmission.

The performance of the system in term of signal to noise ratio E_s/N_o in dB, without equalization is given in Table 1 for different priority streams (P) and for different code rates. The HP stream (H) is simulated in a Rayleigh fading channel. However, the MP and LP streams (M, L) are simulated in a Rician fading channel (K=10). However, from this table, on can see that with powerful channel coding, the error-floor (EF) due to the high power interference will be highly reduced, which implies the high benefits of channel coding.

A T_g of 60-50 μsec for this channel using the more powerful channel coding (the second raw for each priority stream) is a good choice. This leads to a data rate of about 5 Mbps, 7.5 Mbps and 7.5 Mbps for the HP-, MP- and LP-Streams, respectively.

6 Conclusions

We have analyzed the effects of echoes exceeding the guard-time. Performance rapidly degrades as echoes slightly exceeding the guard-time. This will produce an error floor in the bit error rate curves of the uncoded or coded system.

Two methods for compensating this loss may be applied: sub-channel equalization or powerful channel coding. The positive effects of equalization increase as the number of taps increases but the complexity and the adaptation times increase as well. On the other hand the loss due to the short guard interval can be compensated by using a more effective channel coding, with simple channel estimation which can reduce the bit error rates to the desired values.

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