OFDM WITH GUARD INTERVAL AND SUB-CHANNEL EQUALIZATIONIN A 2-RESOLUTION TRANSMISSION SCHEME FOR DIGITAL TELEVISION BROADCASTING

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ABSTRACT *The interest toward a new digital terrestrial TV system, that allows improved quality and receiver portability, leads to the choice of a modulation system suitable to operate in a very hostile environment (interferences, multipath). A very effective solution, already experimented with Digital Audio Broadcasting, is the orthogonal frequency division multiplexing (OFDM) technique with a guard interval. Unluckily such interval grows from few ten micro seconds, sufficient for typical urban propagation environment, up to 600 micro seconds, when the emerging Single Frequency Network (SFN) is considered. A long guard interval (as required in a SFN) results in a excessive prolongation of the symbol period (T) and thus in bandwidth efficiency loss. In addition, the use of a too long symbol period is limited by the effects of time selectivity of the channel and by technological limitations. In this light, in this paper is suggested the use of a shorter guard interval and a low complexity equaliser at the output of each subcarrier to reduce the error produced by the echoes exceeding the guard interval duration.*

A multiresolution (MR) transmission scheme has been adopted in order to eliminate the sharp threshold effect in the fringes of the broadcast area. The system performances of a 2-resolution OFDM system have been evaluated by computer simulation on a frequency-selective channel for Rice and Rayleigh propagation environment.

1. Introduction

The last decade has seen an increasing interest in the possibility of digital terrestrial television broadcasting (dTTb), allowed by the developments in the methods for coding of television pictures from a source rate of over 200Mbit/s to very low bit-rates. In particular, today, it appears feasible to receive high picture quality at home at about 12Mbit/s (fixed receivers with directive antenna). In addition, lately, this interest doesn't neglect the possibility to achieve different services with lower picture quality (i.e. portable or mobile television with set-top antenna), and consequently lower data rate. Besides this, the need of an efficient use of the available spectrum resources suggests the planning (see European dTTb project), in the long term, of a Single Frequency Network (SFN)*. In this light, there are two main problems that a new digital transmission scheme has to face. The first problem, particularly heavy in urban areas, is the multipath propagation which produces intersymbol interference (ISI) when the delay spread is in the same order of magnitude as the symbol period. This problem becomes particularly serious for a portable receiver due to the absence of a direct path (Rayleigh channel). In addition, in the case of a SFN, the echoes may have delays up to some hundred of micro seconds, thus making much more critical the propagation environment.

Besides this, the digital signal has to operate in a very hostile interference environment. In fact, in a first phase the dTTb service will have to share the band with existing broadcast analog television. All these factors lead to a suitable choice of a modulation scheme to overcome these problems. A very attractive modulation method that can be used to reduce the effects of interference and multipath is called orthogonal frequency division multiplexing (OFDM). The use of a guard interval in an OFDM system is a very simple solution to overcome the problems of multipath as far as the channel echo durations are smaller than the guard interval length. The main drawback in the use of a guard interval is that it results in a prolongation of the symbol period at the expense of bandwidth efficiency. Furthermore, an excessive prolongation of T is limited by the effects of time selectivity of the channel and by technological limitations (number of required samples of the FFT, non-linearities in the amplifier, mixer stages and oscillator instabilities). For this reason in this paper the possibility to improve bandwidth efficiency by allowing some echoes to exceed the guard interval is investigated. The consequent performance degradation is restored by using a two taps equalization on each sub-carrier.

The second problem is that a traditional single-resolution transmission technique suffer from a sharp threshold effect in the fringes of the broadcast area. Thus a receiver located at grater distance (or at bad reception conditions) can not receive a full quality image or may risk receiving nothing. However, by using a multiresolution modulation (MRM), it would be at least possible to receive a reduced quality image.

The aim of this paper is to evaluate the performance of a 2-resolution OFDM transmission system, that combines guard interval with a simple equalization on each sub-carrier, under a frequency selective channel. The paper is organised as follows: in section 2 we recapitulate the main characteristics of the terrestrial transmission channel. In section 3 the expression of the OFDM signal as a function of the guard interval is analytically derived. In section 4 we briefly examine the principle of MRM by giving an example of 16-Quadrature-Amplitude-Multiresolution-Modulation (16-QAMRM). In section 5 a Multiresolution OFDM system which combines guard interval and sub-channel equalization is considered and its performance and simulation results are discussed. Finally, section 6 is devoted to some conclusions and remarks.

2. Channel model

In continuous channel modelling, the channel is assumed to be WSSUS (Wide Sense Stationary Uncorrelated Scatters) and characterised by its delay power spectrum $P(\tau)$ [1,2] defined as

$$R_h(\tau,\tau') = < h(\tau)h^*(\tau') > = P(\tau)\delta(\tau-\tau') \qquad (2.1)$$

where $h(\tau)$ is the equivalent low-pass impulse response of the channel, δ is the Dirac delta function, and the averaging is

* A single frequency Network is a transmitter network with overlapping coverage area, transmitting the same program in the same frequency channel at the same instant.

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done over the profiles. Each impulse response is a sample of the channel between a particular location of the transmitter and the receiver at the time of measurement. For fixed and portable reception a quasi-static channel model can be considered, that is, the channel is assumed time-invariant over many symbol periods [3,4].

In the discrete channel model (convenient for channel measurement and computer simulation), the complex envelope of the channel impulse response at a given time is represented by [5]

$$h(\tau) = \sum_{s=0}^{L} \rho_s \, \delta(\tau - \tau_s) \, e^{j\varphi_s} \tag{2.2}$$

That is, the transmitted impulse $\delta(\tau)$ is received as the sum of L+1 paths. In the case of fixed reception (roof top antenna) the received signal is characterised by a strong specular signal component (due to the line of sight) and a number of scattered components (echoes due to multipath propagation) with some delay respect to the specular component. Such a transmission channel can thus be modelled as a Rician fading channel, where the Ricean factor K is the ratio between the power of the specular component and the power of the scattered components. When a portable receiver is used the specular component disappears; because of this, the Ricean factor K becomes 0 and the channel turns into a Rayleigh one. To model this situation, path number 0 in expression (2.2) represents the line of sight. Thus, in presence of line of sight, parameter ρ_0 will be modelled as a Ricean variable with Ricean factor K; the amplitudes ρ_s of the scattered paths (s=[1,L]) will be distributed according to Rayleigh probability density function with second order moment $E(\rho^2{}_s)=P(\tau_s)$ and attenuated by the K-factor. The phases ϕ_s are random variables uniformly distributed in $(-\pi,\pi)$. For what concerns the arrival times, a first-order approximation assumes that objects which cause reflections and refractions in an urban area are located randomly in space, giving rise to a Poisson distribution of the *arrival times* with some mean arrival rate σ [5]. More precisely, if we let t_0 be the line-of-sight delay, one may conjecture that the sequence of excess delays $\{\tau_k=t_k-t_0;\ k=1,\ldots,\infty\}$ of the retarded paths is a Poisson sequence on $(0,+\infty)$. Thus, according to this model, the inter-arrival time between the s-th path and the (s-1)-th path $(\tau_s-\tau_{s-1})$, is described by the exponential probability density function

$$P_T(\tau_s / \tau_{s-1}) = \begin{cases} \sigma \exp[-\sigma(\tau_s - \tau_{s-1})] & \tau_s - \tau_{s-1} > 0 \\ 0 & \tau_s - \tau_{s-1} \le 0 \end{cases}$$

The expected value of this distribution is equal to $1/\sigma$ which represents the average inter-arrival time. For a discrete model represented by $P(\tau)$ and σ, the rms multipath spread for a subclass of profiles is given by

$$T_m(\sigma) = \left[\frac{\sum_k (\tau_k - D)^2 P(\tau_k)}{\sum_k P(\tau_k)} \right]^{1/2} \qquad D = \frac{\sum_k \tau_k P(\tau_k)}{\sum_k P(\tau_k)} \tag{2.3}$$

where τ_k is a random variable representing Poisson arrivals with mean rate σ. A typical delay profile can be expressed by an exponential function:

$$p(\tau) = \begin{cases} \dfrac{1}{T_m} e^{-\tau/T_m} & for\ \tau \ge 0 \\ 0 & for\ \tau < 0 \end{cases} \tag{2.4}$$

where τ represents the excess delay respect to the first ray; T_m is the delay spread, and it is a measure of the width of $p(\tau)$. Typical values are T_m = 0.7 µs for rural terrain, T_m = 7 µs for urban area and up to T_m = 100 µs for mountainous terrain [6]. The ratio $T_r=T_m/T$ (normalised delay spread), where T is the symbol period, is a parameter that describes the severity of the frequency-selective fading. The extension of this model to a SFN leads to a cluster distribution of echoes in (2.2); each cluster, due to a different transmitter, is characterised by an exponential decaying profile and is attenuated according to the distance from the receiver.

3. The OFDM technique

OFDM modulation system, considered in the following, consists of a large number of carriers, equally spaced in frequency, with each carrier digitally modulated. The spectrum of each modulated carrier is allowed to overlap the spectrum of its adjacent carriers [7,8]. The orthogonality is due to the fact that the spectrum of each modulated carrier has a sinx/x shape and the distance between the subcarriers is equal to the inverse of the symbol period T (see Fig.1).

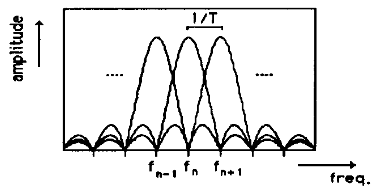

Fig.1 Power density spectra of subchannel components

Thus, a first characteristic of this scheme is the spectral efficiency; in fact, with the increase of the number of the subcarriers and, hence, of the symbol period, the spectrum of an OFDM signal tends asymptotically towards an ideal rectangular spectrum. For example, in the case of a 4 PSK modulation of the single carriers, the asymptotic spectral efficiency is 2 bits/Hertz and is equivalent to a single carrier 4 PSK system with roll-off coefficient equal to zero.

The frequency division is achieved, not by band pass filtering, but by baseband processing. The big advantage of this approach is that both transmitter and receiver can be implemented using efficient Fast Fourier Transform (FFT) techniques. In addition, the OFDM signal gives excellent performance in the presence of multipath because the symbol period of each carrier is much longer than the delay spread of typical multipath reflections.

However, in presence of particularly bad channel conditions, the inter-symbol interference affects the properties of orthogonality between the modulated carriers. As, due to the temporal coherence of the channel and technological limitations, is not possible to increase the number of carriers beyond a certain limit, a very simple solution is the use of a guard interval [9,10] The general principle of the multicarrier concept is shown in fig.2, where g(t) and g'(t) are respectively the transmitter and receiver filter, $\omega_n = n\Delta\omega$ (n= -N/2,...,0,...,N/2) denotes the n-th carrier frequency and $a_n(i)$ is a complex number representing the emitted symbol of the n-th channel at time instant iT.

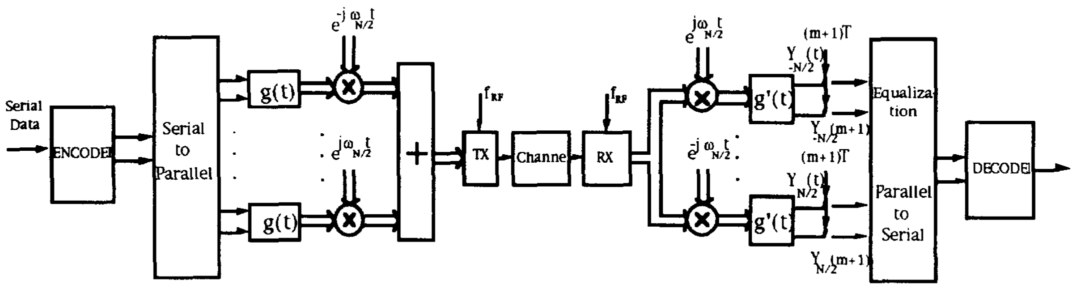

Fig.2 Multicarrier system

In the case of OFDM modulation with guard interval, the transmitting and receiving filters are shown in fig.3 where T is the duration of the transmitted impulses and T_g is the "guard interval". The carrier-frequency distance is $\Delta f= 1/(T-T_g)$.

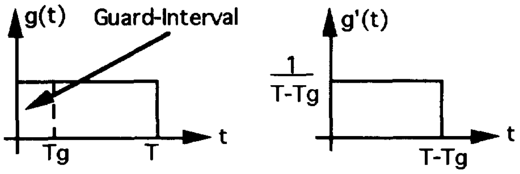

Fig.3 Transmitting and receiving filters

Assuming ideal channel conditions, the received multicarrier signal at the input of the receiver filter-bank can be expressed as

$$x(t)=\sum_{n=-N/2}^{N/2}\sum_{i=-\infty}^{\infty}a_n(i)g(t-iT)e^{j\omega_n t} \qquad (3.1)$$

The demodulated signal, at the output of the k-th receiver filter, is

$$y_k(t)=\left(x(t)e^{-j\omega_k t}\right)*g'(t) \qquad k=-N/2,\ldots 0,\ldots N/2$$

After the calculation of the convolution we obtain

$$y_k(t)=\sum_{n=-N/2}^{N/2}\sum_{i=-\infty}^{\infty}a_n(i)e^{j\Delta\omega_{n,k}iT}r_{n,k}(t-iT) \qquad (3.2)$$

where $\qquad \Delta\omega_{n,k}=(n-k)\Delta\omega$

$$r_{n,k}(t)=\frac{e^{j\Delta\omega_{n,k}t}-1}{j(T-T_g)\Delta\omega_{n,k}} \qquad 0\le t<T-T_g \qquad (3.3a)$$

$$r_{n,k}(t)=\frac{e^{j\Delta\omega_{n,k}t}-e^{j\Delta\omega_{n,k}[t-(T-T_g)]}}{j(T-T_g)\Delta\omega_{n,k}} \qquad T-T_g\le t<T \qquad (3.3b)$$

$$r_{n,k}(t)=\frac{e^{j\Delta\omega_{n,k}T}-e^{j\Delta\omega_{n,k}[t-(T-T_g)]}}{j(T-T_g)\Delta\omega_{n,k}} \qquad T\le t<2T-T_g \qquad (3.3c)$$

$$r_{n,k}(t)=0 \qquad elsewhere \qquad (3.3d)$$

Sampling equation (3.2) at time t = (m + 1)T we obtain:

$$y_k[(m+1)T]=a_k(m)r_{k,k}(T)+\sum_{i\ne m}a_k(i)r_{k,k}[(m-i+1)T]+$$

$$\sum_{n\ne k}\sum_{i}a_n(i)e^{j\Delta\omega_{n,k}iT}r_{n,k}[(m-i+1)T] \qquad (3.4)$$

where the second term at the right-hand side of (3.4) represents *intersymbol interference* (ISI), due to symbols in the same channel (n=k), and the third term represents *adjacent channel interference* (ACI), due to symbols in different sub-channels. The requirement of zero ISI and ACI is fulfilled, in fact considering the equations (3a,b,c,d) it is easy to demonstrate that

$$r_{k,k}[(l+1)T]=0 \qquad\qquad l=\pm1,\pm2,..$$
$$r_{n,k}[(l+1)T]=0 \qquad n\ne k \qquad l=0,\pm1,\pm2,..$$

furthermore, as $r_{k,k}(T)=1$, it follows that $y_k[(m+1)T]=a_k(m)$. In addition, due to the presence of a guard interval, the zero ISI and ACI condition is satisfied in the finite time interval $T-T_g<t<T$, as shown in the eye-pattern of fig.4.

Fig.4 OFDM eye-diagram (ideal channel, T_g=0.2T)

Digital implementation of the OFDM modem

Let's consider expression (3.1) in the time interval [iT,(i+1)T]:

$$x(t)=\sum_{n=-N/2}^{N/2-1}a_n(i)e^{j\omega_n t} \qquad (3.5)$$

In the following we assume to adopt an even number N of sub-carriers. For a large number of sub-carriers, the effect of the side-lobes of the OFDM signal can be neglected, thus we can consider the signal as band limited. By the sampling theorem we must sample (3.5) at the minimum rate $\Delta t=(T-Tg)/N$. By evaluating (3.5) at the sampling instant $m\Delta t$ and substituting the index n with $n'=n+N/2$ we get

$$x(m\Delta t)=x_m=(-1)^m\sum_{n'=0}^{N-1}a_{n'}(i)e^{j\frac{2\pi}{N}mn'} \qquad m\in[0,\ N+M-1] \qquad (3.6)$$

where N+M=T/Δt. Thus, apart from an irrelevant constant factor, the term on the right-hand side of (3.6) is the Inverse

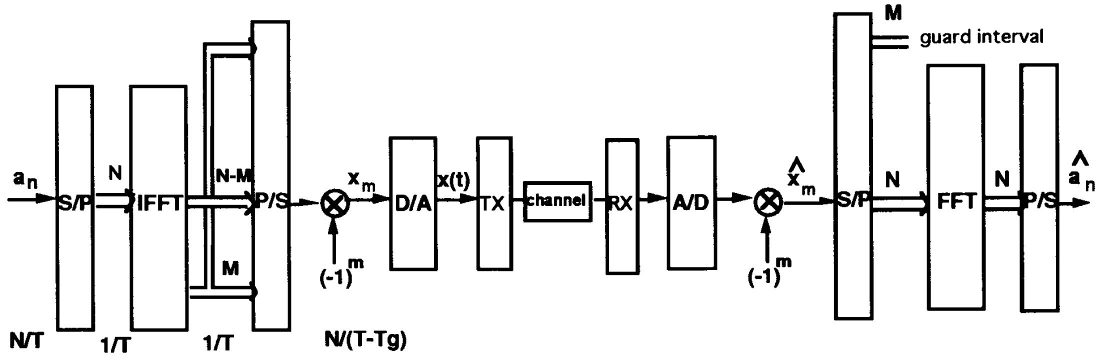

Fig.5 Digital implementation of OFDM system

Fourier Transform (IFFT) of the transmitted symbol sequence; the number of samples used for the IFFT is N. Since every one of the carriers has an integer number of cycles in the integration period $T\text{-}T_g$, this extending of the baseband pulses to a period T can be easily implemented by cyclically prefixing the block of N samples of the output of the IFFT by a repeat of the last M samples. The integration in reception (i.e., collection of the samples for the FFT) is done over only the latter $T\text{-}T_g$ of each symbol, when the transient response of the channel is assumed to have subsided to a negligible level (see fig.5 where for simplicity the circuits for the recovery of carrier and symbol timing have been omitted).

To take into account the effect of a frequency-selective channel, we may express the impulse response by eq. (2.2). The signal at the input of the filter-bank is $\tilde{y}(t) = x(t) * h(t)$ with x(t) expressed by (3.1). After some calculations, it is possible to obtain the following expression for the demodulated signal:

$$\tilde{y}_k(t) = \sum_{n=-N/2}^{N/2} \sum_{i=-\infty}^{\infty} \sum_s h_s e^{-j\omega_k \tau_s} a_n(i) e^{j\Delta\omega_{n,k} iT} r_{n,k}(t - \tau_s - iT) \quad (3.7)$$

where $h_s = \rho_s e^{j\varphi_s}$

If we sample (3.7) at time t=(m+1)T and if $\tau_{max}<T_g$ (that is, the memory of the channel is less than the guard interval) then the two terms of ISI and ACI continue to be zero and $r_{k,k}(T-\tau_s)=1$ thus we obtain:

$$\tilde{y}_k\left[(m+1)T\right] = a_k(m) \sum_s h_s e^{-j\omega_k \tau_s} = H(f_k) a_k(m) \quad (3.8)$$

Where H(f) is the Fourier transform of the impulse response $h(\tau)$. Equation (3.8) shows that the effect of the channel consists simply in multiplying the transmitted symbol $a_k(m)$ by a factor corresponding to the frequency channel response at the time instant (m+1)T and at the frequency f_k. Equation (3.8) provides a simple way for channel transfer function estimation, necessary for coherent reception (as we assume in this paper). In fact, if the channel is known, the transmitted information is completely recovered by multiplying the received symbols by $1/H(f_k)$. In practice the channel response $H(f_k)$ can be derived by sending periodically a known test pattern. Let $A_k(m)$ be this test symbol, then the channel response on the k-th sub-channel at the time instant (m+1)T is given by:

$$H(f_k) = \tilde{y}_k[(m+1)T] / A_k(m) \quad (3.9)$$

The frequency of the test pattern depends on how fast the channel varies. It is worth noting that this equalization is perfect as long as the max echo delay is less than the guard interval. The optimal repetition rate of the test symbols could be a point of further investigation.

If the max delay is known, the condition $\tau_{max}<T_g$ can be fulfilled by determining the appropriate number of subcarriers, when the bandwidth efficiency has been fixed. For typical urban propagation environments $\tau_{max} < 30\mu s$ whereas, in the case of Single Frequency Networks (SFN) and omni-directional antennas, echoes with delay of $600\mu s$ may occur. It is evident that, in order to satisfy the condition $\tau_{max} < T_g$, the symbol duration T has to be increased with the maximum delay. However, an excessive prolongation of T is limited by the effects of time selectivity of the channel and by technological limitations (number of required samples of the FFT, non-linearities in the amplifier, mixer stages and oscillator instabilities). Besides this, this protection is achieved at the expense of bandwidth efficiency as the carrier distance is $\Delta f = 1/(T-T_g) > 1/T$. In order to limit these drawbacks due to the use of a too long guard interval, we suggest to replace the simple multiplication (3.9) with a low complexity equaliser at the output of each subcarrier to reduce the error produced by the echoes exceeding the guard interval duration (see fig.6). If we consider the interfering term in eq (3.7), we note that, for the practical values of the expected delays and of T, the ISI term will only be caused by one preceding symbol. Since the number of taps in the equalizer delay line is chosen so that the total delay is approximately equal to the multipath delay spread introduced by the channel, we suggest the use of a two taps equaliser with weight coefficients $\{c_j\}$ adjusted in order to minimise the mean-square value of the error (MSE criterion)

$$\varepsilon_k(m) = a_k(m) - \hat{a}_k(m)$$

where $a_k(m)$ is the transmitted symbol in the m-th signalling interval on the k-th sub-channel, and $\hat{a}_k(m)$ is the estimate of the symbol at the output of the equaliser defined by

$$\hat{a}_k(m) = c_{0,k} \tilde{y}_k[(m+1)T] + c_{1,k} \tilde{y}_k[mT]$$

Thus, the function that has to be minimised at the output of each equalizer in order to compute the taps is

$$E\left\{\left|a_k(m) - \left(c_{0,k}\tilde{y}_k[(m+1)T] + c_{1,k}\tilde{y}_k[mT]\right)\right|^2\right\} \tag{3.10}$$

The simplest adaptation algorithm is the stochastic gradient algorithm which converges to the optimal solution of (3.10). The adaptation equations for the tap coefficients are

$$c_{p,k}^{(m+1)} = c_{p,k}^{(m)} - \alpha_k[\hat{a}_k(m) - a_k(m)]\tilde{y}_k^*[(m+1-p)T]$$
$$p = 0,1 \quad k = -N/2 \ldots N/2$$

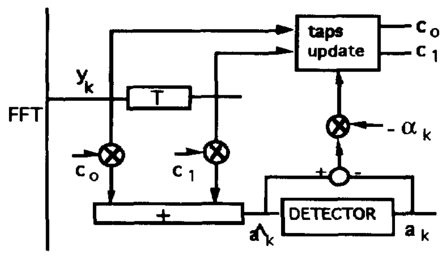

Fig.6 *Two tap equalizer*

where α_k is the step size coefficient on the k-th sub-channel. Two examples of the manner in which the MSE coefficients adapt to a sudden change in channel conditions are presented in Figs.7. The left-hand frames (fig.7a) show the initial convergence characteristics of the MSE algorithm with different step sizes and a fixed channel. We observe that by selecting α_k=0.2 we obtain relatively fast initial convergence without any relevant increase in the excess mean square error. The right-hand frames (fig.7b) show the error magnitude time histories for different normalized channel delay-spreads (T_r=T_m/T) and for a fixed step size value (α_k=0.2).

In the above discussion it was assumed that the receiver has knowledge of the transmitted information sequence in forming the error signal between the desired symbol and its estimate. A practical scheme for continuous adjustment of the tap weights may be either a decision-directed mode of operation in which decisions on the information symbol are assumed to be correct and used in place of $a_k(m)$ in forming the error signal $\varepsilon_k(m)$, or a known test pattern sequence may be inserted in the information-bearing signal either additively or by interleaving in time. In the decision-directed mode of operation, the error signal becomes ε_k=$\tilde{a}_k$-$\hat{a}_k$ where $\tilde{a}_k$ is the decision of the receiver based on the estimate $\hat{a}_k$. It is worth noticing that such an equaliser could possibly be time-shared by some sub-channels, thus decreasing the receiver complexity.

4. Multi-resolution transmission

The principle of multiresolution is to match different quality levels to different channel capacities [13,14]. In this way the receiver closer to the emitter can decode the full quality signal while the distant receiver has access to the lower-resolution quality, thus providing a stepwise graceful degradation. This idea can be carried out in the modulation domain. In this case each constellation consists of "clouds" of miniconstellations where the detail information is carried by the points in the clouds, and the " coarse" information is carried in the clouds.

The receiver first decodes the likeliest cloud (coarse information), "subtracts" the decoded cloud value from the

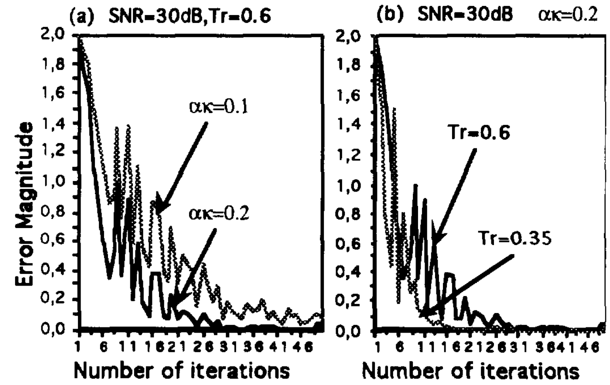

Fig.7a,b *Convergence of MSE error for different step size coefficients (a); Convergence of MSE error for different normalized channel delay-spread Tr(b)*

received point, and then decodes the likeliest point within the cloud (detail information). In fig.8 is displayed a MR-16 QAM constellation. For every four composite bits, two bits select one of the four clouds while the remaining two bits select one of the four "satellites" within the selected cloud. The relative distances between intercloud constellation points (d_1) and intercloud points (d_2), whose ratio is a design parameter λ=d_2/d_1, must be optimised with respect to the degree of desired protection for the coarse information (2bits).

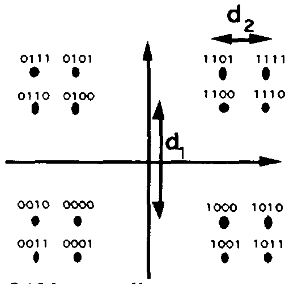

Fig.8 *MR 16 QAM constellation*

For an AWGN channel the symbol error rate (SER) of the first resolution bits (the most significant bits) can be derived easily:

$$SER \cong \frac{1}{2}\left[erfc\left(\frac{d_1/2}{\sqrt{2N_o}}\right) + erfc\left(\frac{d_1(1+2\lambda)/2}{\sqrt{2N_o}}\right)\right]$$

where No/2 is the Gaussian noise spectral density. In the same manner the SER for the second resolution bits can be computed as follows:

$$SER \cong \frac{1}{2}erfc\left(\frac{d_1/2}{\sqrt{2N_o}}\right) + erfc\left(\frac{\lambda d_1/2}{\sqrt{2N_o}}\right)$$

5. Performance evaluation

The 2-resolution OFDM system shown in fig.9 has been at first simulated without equalizer on Rayleigh frequency-selective fading channel

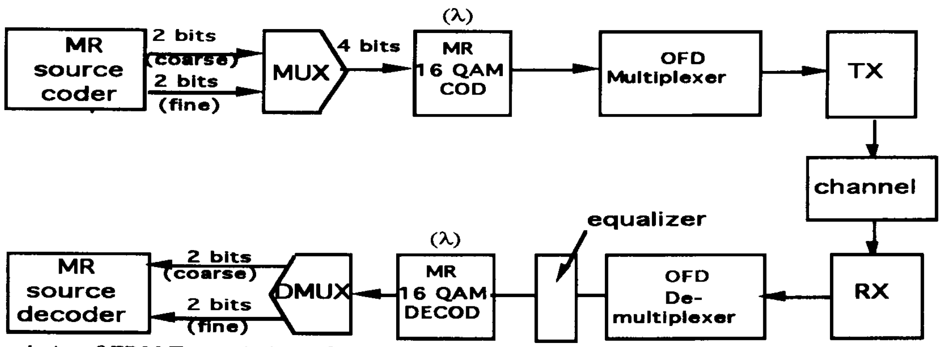

Fig.9 *2-resolution OFDM Transmission scheme*

The Bit Error Rate (BER) performances of the "coarse" information have been reported in fig.10 vs. the average signal to noise ratio E_b/N_0 (E_b is the average bit energy). The channel impulse response has been approximated by the exponential profile reported in equation (2.4). The multirisolution constellation parameter λ has been set to 0.5.

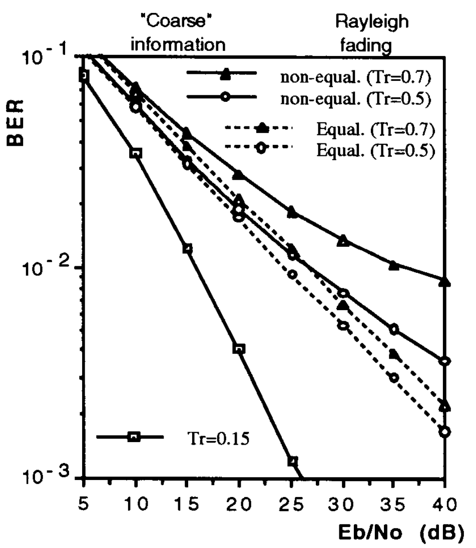

Fig.10 *Performances of the "coarse" resolution on a Rayleigh frequency-selective fading channel with (dotted line) and without (continual line) equalizer*

For a fixed environment, the evaluation of the performances has been carried out by computing the impulse response for a large number of channel events (i.e. in a large number of receiving sites) and hence evaluating the instantaneous values of the error probability using a "semi-analytic" approach. A guard interval T_g = T/5 has been assumed and no channel coding has been introduced. The effect of delay spread on the signal has been characterised by the parameter $T_r = T_m/T$. The existence of a minimum limit of the error-probability for the system without equalizer (continual lines), corresponding to the asymptotic tendencies of the curves, results from the intersymbol and adjacent channels interference when the maximum delay exceeds the guard interval. The greater the ratio T_m/T, the higher is the minimum error-probability; when T becomes large compared with T_m, the error probability curve tends to that of the non-selective Rayleigh model.

To determine the effectiveness of the MSE equaliser in reducing ISI, the equalized OFDM system (dotted line) has been subsequently simulated on Rayleigh fading channel and compared with the non-equalized one. The reported curves show how the two tap equalizer strongly reduces the error floor. The effect of the equalizer in reducing ISI is also pointed out in figs.11a,b where the received constellation is shown without (fig.11a) and with (fig.11b) equalizer on a frequency-selective Rician channel (K=10dB).

Note that the overall performance of a MR system depends on the respective performance of each resolution. In this light, in fig.12 we reported the simulated Bit Error Rate curves of the "fine" (MR-16 QAM) and "coarse" (QPSK) information with different λ values. In particular, the fine resolution performances were evaluated in presence of a Rician channel (thus simulating the propagation conditions that a fixed receiver with roof-top antenna would experience) while the BER for the coarse information was evaluated in presence of a Rayleigh channel (thus reproducing the impairments that a portable receiver would undergo). As a first intuitive result, these figures point out that by reducing λ the performance of the low resolution system improve (greater distance between "clouds") while the opposite happens for the fine resolution receiver (smaller distance between adjacent points in a "cloud"). Thus, the choice of the design parameter λ must be a proper trade-off between a sufficient protection grade for the portable receiver (that undergoes the worst propagation condition) and the performance of the fine resolution receiver.

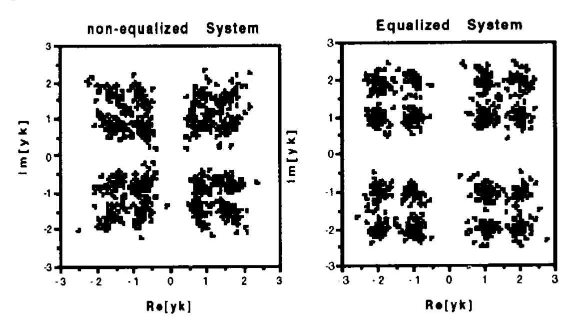

Fig.11a,b *Signal space of OFDM sub-carrier under frequency selective Rice channel with relative delay-spread Tr=0.6 without equalizer (a), with equalizer* (b).

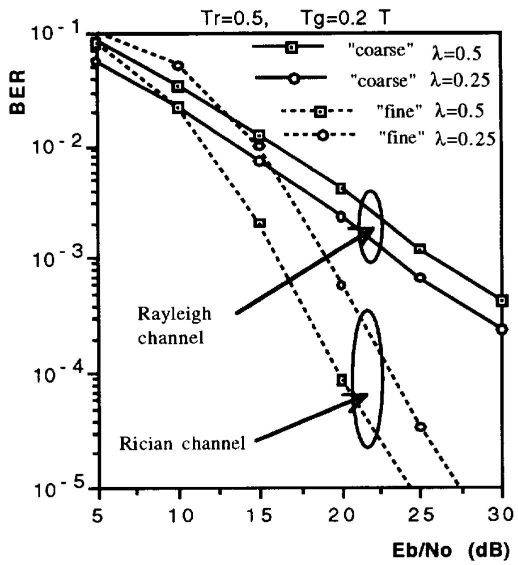

Fig.12 Performances of the OFDM 2-Resolution Transmission Scheme on a Rayleigh Fading Channel and on a Rician Fading Channel (K=10dB) for λ=0.5, λ=0.25

It is worth mentioning that the different antenna gains of the different receivers where not considered. In case of system optimisation, one should take into account the respective gains of different antennas. The use of powerful channel coding (e.g.concatenation of inner convolutional and outer Reed-Solomon) would greatly improve system performances.

6. Conclusions

In this paper it was pointed out that an OFDM technique, combined with guard interval and low complexity sub-channel equalization is an appropriate solution for fixed and portable digital terrestrial television in case of long echo delays (e.g. SFN or mountainous terrain). In fact, as an alternative to increasing the symbol duration, we have analysed the effect of reducing the guard interval below the longest echo delay and contemporaneously introducing a two taps equalization on each sub-channel.

The performances of this system configuration, adopting a 2-resolution QAM signal constellation, have been illustrated by simulation results which are based on realistic multipath radio channel conditions in the case of Ricean and Rayleigh fading. The obtained results clearly show the positive effects of equalization in reducing the error floor due to ISI and ACI.

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