DIGITAL SOUND BROADCASTING TO MOBILE RECEIVERS

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ABSTRACT

Many European countries have shown an increasing interest in the development of a new audio broadcasting service with a view to providing an improved sound quality on portable and mobile receivers.

Digital techniques have progressed over the past few years in the areas of sound programme production and source bit rate reduction as well as in the field of channel coding and modulation. Within the framework of the DAB (Digital Audio Broadcasting) EUREKA 147 project, our research institute has designed a new broadcasting system, fully digital from studio to user.

This paper deals with a promising modulation and channel coding system, suitable for digital broadcasting through the particularly hostile urban radio channel.

A successful field demonstration has been organized under the auspices of the European Broadcasting Union during the WARC-ORB 88 in the town of Geneva, to validate these new concepts of digital sound broadcasting, combining spectrum and power efficiency.

1. INTRODUCTION

From the technical point of view, the development of such a new digital audio broadcasting system is related to a twofold problem :

- it requires the development of a sophisticated source coding system. It is generally admitted that, in this respect, a monophonic sound with a quality equivalent to that produced by compact disks can be reduced to 128 kbit/s or less.
- it also requires the design of a channel modulation and coding system that can be used by mobile receivers up to 1 GHz or more.

This article deals exclusively with this second problem.

For reception in a built-up area, the channel impulse response usually extends over a few microseconds, and in some cases as much as 10 μs or more. This being so, frequency-non-selectivity concerns only low bit-rate transmissions (a few tens of kbit/s) and can under no circumstances constitute a valid hypothesis for high-quality sound broadcasting.

Given the high selectivity of the channel, conventional equalization techniques are very difficult to implement. More sophisticated techniques, such as Viterbi equalization, do not apply because of the excessively high number of states required in the Viterbi decoder.

The methods generally adopted to resolve this type of problem are based on the use of M-orthogonal alphabets [1] [2] which increase the duration of symbols and allow for paths discrimination. Unfortunately intercorrelation constraints between the various alphabet sequences lead to a major increase in the bandwidth which is far from compensated by the increased number of bits transmitted per symbol. The spectral efficiency of this type of system is therefore totally incompatible with the broadcasting constraints.

The system described in this article, combining spectrum and power efficiency, is mainly based on the conjunction of the Orthogonal Frequency Division Multiplexing technique (already proposed for HF data transmission in an ionospheric channel or for transmissions through telephone networks [3] [4]), and a coding strategy associated with diversity in the frequency domain.

Section 2 gives a general overview of the urban mobile radio channel and its effects on data transmission. The general principles of the COFDM (Coded Orthogonal Frequency Division Multiplex) are explained in section 3. Section 4 gives a detailed representation of the signals and the decoding procedure. The performances of the system in the so-called selective Rayleigh channel are then presented in section 5. Section 6 describes the behaviour of the system when the conditions of temporal and frequential coherence of the channel are not fully met (very high speed reception and anormally extended channel delay spread). In section 7, the use of such a system in a broadcasting network is discussed. The realization aspects are finally presented in section 8.

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2. DIGITAL TRANSMISSION IN THE URBAN RADIOMOBILE CHANNEL

The broadcasting channel used for transmission to mobile receivers in a built-up area is a particularly hostile transmission environment. Industrial interference, and mainly multipath propagation caused by natural obstacles, necessitate the implementation of sophisticated modulation devices in order to provide an excellent-quality transmission. One of the major difficulties results from the constant evolution of the channel characteristics as the receiver is moving.

Theoretical studies summarized in CCIR [5] reports established a twofold channel model :

- the first part estimates the value of the mean energy received in small areas (a few hundred wavelengths) ;
- the second part takes into account the combination of various paths originating from discrete reflections and received after scattering on objects (such as trees, other vehicles, etc...) that cannot be considered as mere reflective surfaces.

With regard to the first part of this model, results of experimental research in an urban environment showed that the distribution of mean energy passed from one small area to another follows a log-normal law with a mean value related to free-space propagation.

The second part of the model can be represented by the block diagram of figure 1 :

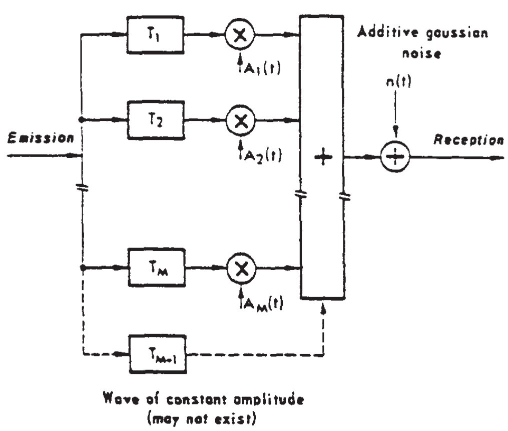

Figure 1

Outline of the transmission channel

The delays T_i originate from the various specular reflections, while the multiplying factors $A_i(t)$ are the consequence of local scattering. If the number of scattered sub-waves is large enough, it can be shown that the modulus of the term $A_i(t)$ follows a Rayleigh distribution. The power spectrum of $A_i(t)$ can be easily deduced from the distribution of the incidence of the sub-waves with respect to the direction of displacement of the vehicle. Assuming this distribution to be uniform and continuous over $[-\pi , \pi]$, the power spectrum of the process $A_i(t)$, translated to baseband and associated with each path, can be written :

$$\gamma A_i(\nu) = \frac{P_i}{\pi\sqrt{(\frac{v}{c}\cdot f_0)^2 - \nu^2}} \quad \text{if } -\frac{v}{c} f_0 < \nu < \frac{v}{c} f_0$$

In this expression, f_0 is the carrier frequency, v is the vehicle speed and c is the velocity of the light.

Moreover, it can be easily understood that the variations of the various Ai(t) are not correlated, because they arise from distinct origins.

Considering the problem of data transmission through this type of channel, a large number of authors [6] [7] [8] pointed out that classical modulation systems (eg 4-PSK or FSK) lead to an irreducible bit error rate, even for large signal-to-noise ratio, caused by two independent phenomena :

- Selectivity, due to the spread of the channel impulse response, will cause intersymbol interference, as soon as the data rate tends to increase.
- Time variation of the channel characteristics, as a result of the changing environment of the receiver, will cause degradation in the phase estimation of the receiver, as soon as the data rate tends to be too low, for a given vehicle speed and carrier frequency.

Even in the case where the channel is neither frequency-selective nor time-selective, the bit error rate as a function of the signal-to-noise ratio is known to decrease very slowly. More than 35 dB of E_b/N_0 is needed for a 4-DPSK modulation scheme, to achieve a 10^{-4} bit error rate (see section 6).

This makes evident the need of sophisticated modulation and channel coding schemes for data broadcasting through the mobile urban channel.

As it has already been said, frequency selectivity must be considered as an unavoidable phenomenon for high quality sound broadcasting. Moreover, the diversity provided by the use of wide-band transmission must be considered as an advantage if the communication system is designed to make use of multipaths rather than be restricted by their presence. Because of the spread of the channel response, it is highly unlikely that fading would simultaneously affect a frequency band covering a few megahertz.

The system described below exploits this property. In addition, it offers an excellent spectral efficiency, which is essential in sound broadcasting.

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This being so, a certain number of sound programmes (e.g. 12 to 16) have to be multiplexed, forming a signal in which each basic data source will benefit from the "wide-band" character of the transmission.

3. GENERAL PRINCIPLES OF THE COFDM SYSTEM

Taking into account the above channel modelling, it is possible to represent the effects of the transmission by combining the channel frequency response and time variation (fig.2). This two-dimensional function characterizes the so-called "selective Rayleigh channel", and admits a decomposition in surfaces of different sizes :

- the small surfaces represent the frequency-time areas where the channel can be considered as locally invariant ;
- the large surfaces indicate the minimum separation for which two small surfaces are statistically independent.

Considering where the channel is invariant on the one hand and statistically independent on the other, this decomposition constitutes the basis of the channel modulation and coding method described in the following sections [9].

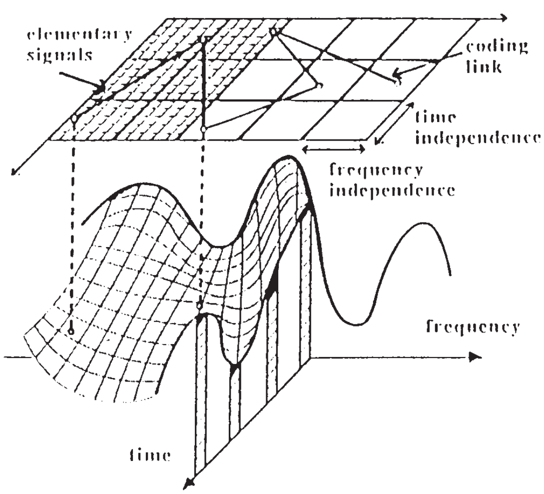

Figure 2

channel frequency response

3.1 OFDM modulation technique

The first principle consists in splitting the information to be transmitted into a large number N of modulated carriers with a low bit rate, in order to reduce the effect of the channel frequency selectivity. The OFDM technique permits the time-frequency domain to be split into small surfaces (fig. 2) with different dimension T_S and $1/T_S$ on the time and the frequency axes respectively.

The modulated OFDM signal can be expressed on an orthogonal base of elementary signals with complex coefficients, where each elementary signal is defined as one of the N emitted carriers during a symbol time T_S, and each coefficient, taking its value from a finite alphabet, represents the modulation applied to each elementary signal.

The main difference between conventional Frequency Division Multiplexing and this technique, is that the spectrum of the different carriers mutually overlap, giving therefore an optimum spectrum efficiency (asymptotically 2 bit/s/Hz for a 4-PSK modulation of each carrier). Nevertheless, the signal verifies orthogonality conditions, so that it is possible to extract the information modulating each carrier without suffering interference due to the presence of the other carriers. Morever, the modulation and the demodulation processes can be undertaken by means of Fast Fourier Transform algorithms.

Regarding the multiple paths of the transmission channel, the condition of perfect orthogonality between carriers is no longer maintained at the receiver input, due to residual intersymbol interference.

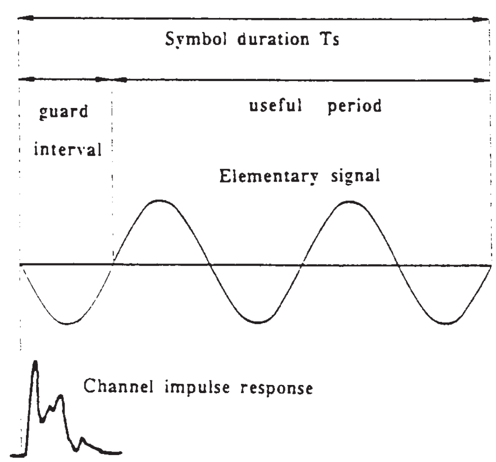

Figure 3

Use of a guard interval to suppress the intersymbol interference

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An asymptotic solution to solve this problem would consist in increasing indefinitely the number of carriers and consequently the symbol duration. But this method is irrealistic taking into account the limitations dictated by the time coherence (see section 6) of the channel (Doppler effect). The retained solution comprises the addition of a safeguard interval before each useful symbol. If the safeguard interval duration is chosen sufficiently longer than the spread of the impulse response of the channel, the useful period of the signal remains free of intersymbol interference and the orthogonality remains perfect (fig. 3).

3.2 Coding scheme

Although the OFDM technique resolves the problem of transmission channel selectivity, it does not suppress fading. The amplitude of each carrier is usually affected by a Rayleigh law (or a Rice law when there is a direct path). This is why an efficient channel coding system is essential.

The second principle of the COFDM system consists on linking by a coding procedure, elementary signals (small squares) transmitted at distant locations of the time-frequency domain (fig.2). This is achieved by convolutional coding (associated to soft decision Viterbi decoding) in conjunction with frequency and time interleaving. The interleaving depth is relative to the dimensions of the large surfaces of the channel representation.

The diversity provided by interleaving plays a vital role in this system. The Viterbi decoder cannot function correctly unless successive samples presented at its input are affected by independent Rayleigh laws. In practice, the distortions to which these samples are subjected have a strong time/frequency correlation. When the receiver is not moving, the diversity in the frequency domain is sufficient to ensure that the system functions correctly. Due to the spread of the channel response (a few microseconds), flat fadings over a few megahertz are very unlikely. From this point of view, the existence of multipaths is a form of diversity and should be considered as an advantage.

4. SIGNALS REPRESENTATION AND DECODING PROCESS

The elementary modulation symbol transmitted during the time interval $T_s = t_s + \Delta$ is given by :

$$x(t) = \sum_{k=o}^{N-1} \mathrm{Re}\left(C_k e^{2i\pi f_k t}\right) \qquad t \in [0, T_s]$$

with $f_k = f_o + k/t_s$

where the equation parameters are the following :

t_s : useful symbol duration, on which the demodulation will be processed.

Δ : guard interval duration.

f_o : transmitting frequency.

N : number of carriers of the multiplex.

C_k : complex element of the modulation alphabet.

The emitted signal is constituted by the juxtaposition in time of the elementary symbols defined above.

The transmitted message fixes the values of the modulation elements C_k, belonging to an alphabet that specifies the type of modulation.

For example, the alphabet corresponding to a 4-PSK modulation is the following :

$$\{1+i, 1-i, -1+i, -1-i\}$$

On the assumption that the guard interval is longer than the channel impulse response, and that the channel varies slowly as compared to the symbol duration (invariance of the channel on small surfaces), the elementary symbol received in the time interval free of intersymbol interference, can be expressed by :

$$y(t) = \sum_{k=o}^{N-1} \mathrm{Re}\left(H_k C_k e^{2i\pi f_k t}\right)$$

where $H_k = \rho_k e^{i\varphi_k}$ is the channel response at the frequency f_k. The signal is translated in baseband by means of a local oscillator at the frequency $f_o + 1/2\,T$ where $T = t_s/N$. The corresponding complex signal thus obtained and sampled at the rate 1/T can be written :

$$z(nT) = (-1)^n . \sum_{k=o}^{N-1} H_k C_k . e^{2i\pi\frac{nk}{N}} \qquad (n = 0 \text{ to } N-1)$$

It appears that the set $\{ H_k C_k : k = \{0,..., N-1\} \}$ is the Discrete Fourier Transform of the set :

$$\{ (-1)^n z(nT)/N : n = \{0,..., N-1\} \}.$$

In fact, the FFT realizes in the discrete-time domain, the equivalent of a bank of N filters, each of them being matched to a given carrier of the multiplex.

The representation of the equivalent receiver using correlation is given in figure 4. The conditions of orthogonality mentioned above clearly appear in the fact that, at the input of the integrator of the row l, the carriers f_k ($k \neq l$) exhibit an integer number of periods $|k - l|$ during the integration time.

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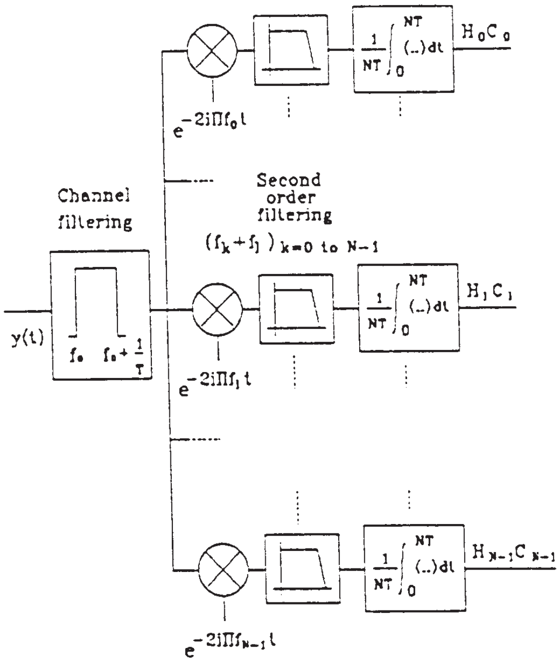

Figure 4

Theoretically equivalent receiver using a bank of N correlators

In absence of noise, the emitted symbols can be recognized without error, without taking into account the problem of H_k estimation that will be examined later. The phenomena of residual bit error rate due to the frequency selectivity have disappeared. Nevertheless, the decrease of the bit error rate as a function of the ratio Eb/No is extremely low in a Rayleigh channel. As previously said, this property justifies the use of an efficient channel coding system.

By introducing the temporal dimension (index j) and the noise, the Fourier Transform process returns the samples :

$$Y_{j,k} = H_{j,k} C_{j,k} + N_{j,k},$$

where $N_{j,k}$ is a complex gaussian noise, each component of which having a variance $\sigma^2_{j,k}$.

The *a posteriori* maximum likelihood criterion consists in maximizing with respect to the elements $\{C_{j,k}\}$ linked by the convolutional coding, the expression :

$$\sum_j \sum_k \| Y_{j,k} - H_{j,k}\, C_{j,k} \|^2 / 2\sigma^2_{j,k}$$

As far as the modulation and demodulation processes of each carrier are concerned, a 4-PSK modulation has been considered at the present time. In the first step of the study, other modulations with a larger number of states (8-PSK, 16-QAM,...) have been discarded, taking into account their reduced power efficiency. Considering the coherent demodulation a 4-PSK-COFDM signal using a convolutional code of rate 1/2, and free distance 10, simulation results point out that a bit error rate of 10^{-3} can be achieved for a mean value of the Eb/No ratio equal to 5 dB in a selective Rayleigh channel. It means that the instantaneous value of this ratio can highly decrease leading to a particularly difficult implementation of the coherent demodulation.

The differential demodulation is an alternative solution of which the essential interest resides in the very great simplicity of its implementation and its absence of inertia after a deep fade. The price to be paid is obviously a degradation in performance which nevertheless remains acceptable and which is in reality minimal if account is taken of the practical limitations of coherent demodulation.

As far as the estimation of the channel frequency response is concerned, differential demodulation consists in using at the time j a simplified estimator of the channel deduced from the time j-1 :

$$H_{j,k} \approx Y_{j-1,k} \,/\, C_{j-1,k}$$

Considering a precoding at the emission following the law,

$$a_{j,k} + i b_{j,k} = (1+i)\, C_{j,k} / C_{j-1,k}\,, \qquad a_{j,k} = \pm 1\,,\ b_{j,k} = \pm 1$$

where $a_{j,k}$ and $b_{j,k}$ constitute the outputs of the convolutional coder, it appears that the corresponding weightings to be used at the level of the Viterbi decoder are :

$$\mathrm{Re}\left(\frac{Y_{j,k}\, Y^*_{j-1,k}}{(1-i)\,\sigma^2_{j,k}}\right) \text{ and } \mathrm{Im}\left(\frac{Y_{j,k}\, Y^*_{j-1,k}}{(1-i)\,\sigma^2_{j,k}}\right)$$

Moreover, these formula indicate that if the signal is affected by a narrow-band interferer, the term $\sigma^2_{j,k}$ in the weightings has the effect of "deleting" the corresponding carriers in the same way as a fading of the same carriers. Taking into account that the spectral analysis of the noise is particularly simple with a COFDM system, this property makes this system extremely attractive for channels which are highly disturbed by noise or industrial interference.

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5. THEORETICAL RESULTS IN A SELECTIVE RAYLEIGH CHANNEL

As far as the performance of the COFDM system is concerned, a wide choice of carrier modulations and channel coding parameters allows a trade-off between bit error rate performance and spectrum efficiency.

For our application, a 4-PSK modulation scheme has been applied to each COFDM carrier, associated to a coherent or differential demodulation.

Additionally, extremely abrupt error-ratio curves may be achieved by using an outer code of Reed Solomon type, concatenated with the inner convolutional code at the price of a slight reduction of spectrum efficiency.

Accordingly, simulation results have been plotted in figure 5 in comparison with the Shannon limit. In particular, for different convolutional code rates lying from 1/4 to 8/9 with constraint length 7, concatenated with Reed-Solomon codes of parameter (n,k), with n = 255 and k = 211 to 243, the spectrum efficiency of the COFDM system lies from 0.4 to 1.5 useful bit/s/Hz and the necessary Eb/No ratio for a transmission without error varies from 3 to 12 dB respectively considering a coherent demodulation.

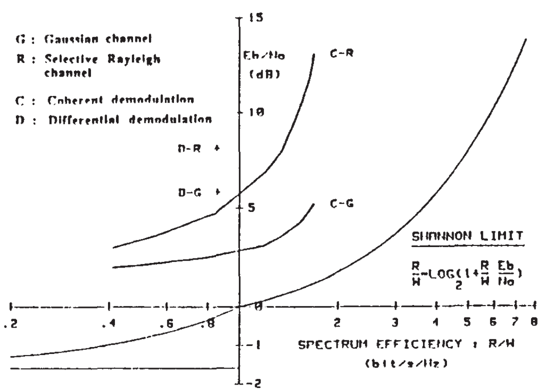

Figure 5

COFDM performances

Although the concatenated codes previously described offer remarkable performance, their implementation in the context of the mass production domestic market is difficult because of the complexity of decoding the Reed-Solomon codes. An alternative solution consists in using CSRS codes (Cyclotomatically Shortened Reed-Solomon) [10] which are, like the Reed-Solomon codes, at a maximum distance separable ($d_{min} = n - k + 1$). The essential interest of these codes is that they do not require processing over the Galois field extension $GF(2^8)$, but only over GF(2). The implementation of the decoding is considerably simplified. The codes that we considered have the form (n , n - 2t), where t is the number of corrected symbols of j bits. For the simulations described hereafter, we have taken j = 12, n = 336 and t = 24.

Accordingly, the performances given in figure 6 are those of a 4-PSK-COFDM system using a convolutional code of rate 1/2, constraint length 7, and free distance 10, eventually concatenated with a CSRS code (336, 288).

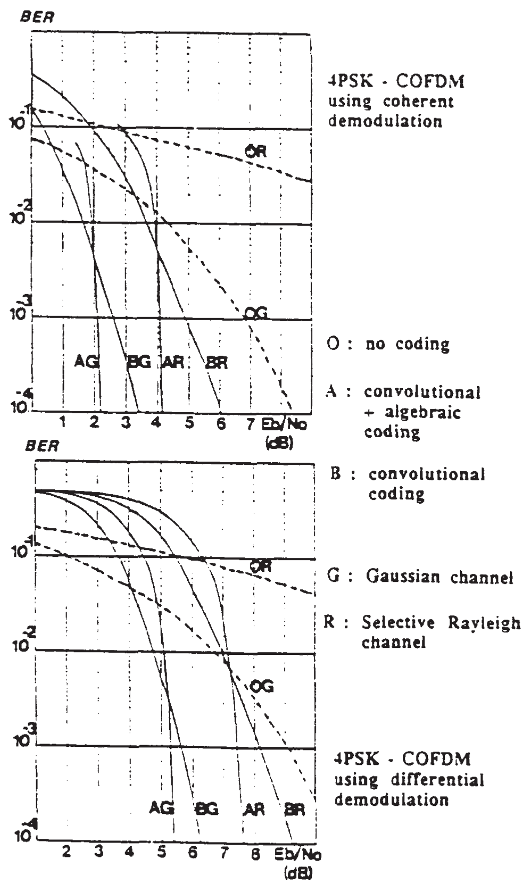

Figure 6

4PSK - COFDM performances

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6. TEMPORAL AND FREQUENTIAL COHERENCE OF THE CHANNEL

This section intends to investigate the time and frequency coherence of the channel in relation to the transmitted signal, in order to determine the COFDM parameters $1/t_s$ and T_s. These notions refer to the small surfaces of the channel representation (section 3) where the channel can be considered as locally invariant.

The channel transfer response can be expressed in the time-frequency domain by :

$$H(f,t) = \sum_i \exp\left(j\,\varphi_i\right) \exp\left(-j2\pi f\tau_i\right) \exp\left(j2\pi f_{di}\,t\right)$$

where φ_i is the phase shift, τ_i the propagation delay and f_{di} the Doppler shift for the i^{th} path.

The channel is then characterized by the scattering function $S(\tau, f_d)$ which represents the power spectrum as a function of the time delay τ and the Doppler frequency f_d [11] [12].

For simulation simplification, the random variables f_d and τ are supposed to be statistically independent. This hypothesis leads to the scattering function of the form :

$$S\left(\tau, f_d\right) = K\; P_\tau(\tau)\; P_{f_d}\left(f_d\right)$$

where $P_\tau(\,)$ and $P_{fd}(\,)$ are the probability density functions of τ and f_d. In most of the channel impulse measurements, the time delay function can be approximated by an exponential distribution $P_\tau(\tau) = \frac{1}{T_m} e^{-\tau/T_m}$ where T_m is the standard deviation of the delays, namely the delay spread (fig. 7).

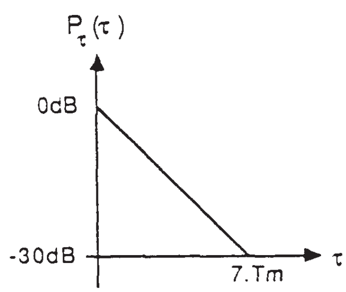

Figure 7

Time delay density function

In addition, by assuming that all angles of incident waves are uniformly distributed, the Doppler function can be given by :

$$P_{fd}\left(f_d\right) = \frac{1}{\pi} \frac{1}{\sqrt{f_{max}^2 - f_d^2}} \quad \text{for } |f_d| < |f_{max}|$$

where f_{max} is the maximum Doppler frequency and is given by $f_{max} = f_0 . v/c$ (v is the speed of the vehicle).

6.1 Frequency-selective fading

The first characteristic of the multipath medium is the frequency variant aspect of the channel transfer function. It may be shown that the reciprocal of the delay spread T_m is a measure of the coherence bandwidth $(\Delta f)_c$ of the channel :

$$(\Delta f)_c \approx 1/T_m$$

The effect of the channel on the transmitted signal is a function of the choice of the signaling interval duration t_s. If t_s satisfies the condition $t_s >> T_m$ the channel introduces a negligible amount of intersymbol interference. This condition implies in the frequency domain that $1/t_s << (\Delta f)_c$ and therefore that the channel is frequency-non-selective with respect to one modulated carrier of the COFDM signal.

The effect of delay spread on the signal can be characterized by the parameter $\mu = T_m/t_s$ and the above condition becomes $\mu << 1$.

Computer simulations have been carried out to analyse the influence of μ on the COFDM signal performance (fig. 8), with and without coding and taking into account a guard interval duration Δ equal to t_s /4. In particular, the results point out the important gain provided by the coding.

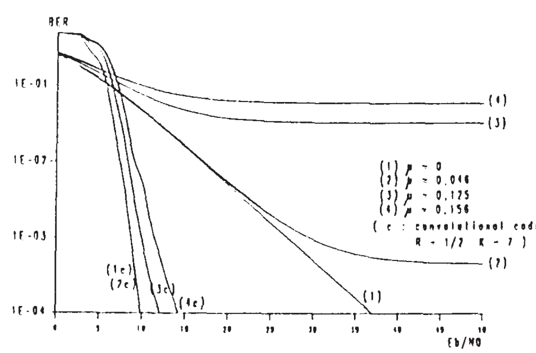

Figure 8

The performance of 4-PSK-COFDM in frequency-selective fading channels (differential demodulation and Δ = ts/4)

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6.2 Time-selective fading

The second characteristic of the multipath medium is the time-variant aspect of the channel transfer function.

The coherence time of the channel can be given by $(\Delta t)_c \approx 1/f_{max}$ where f_{max} is the maximum Doppler frequency.

The channel is called a slowly fading channel, or a time-non-selective channel, if the channel attenuation and phase shift are essentially fixed for the overall duration T_s of the signaling interval.

This condition expressed by $T_s << (\Delta t)_c$ implies that $\beta << 1$ where β is equal to $f_{max}\, T_s$ and characterizes the effect of the rapidity of the fading on the signal.

Simulation results, with and without coding, for different values of β are plotted in figure 9.

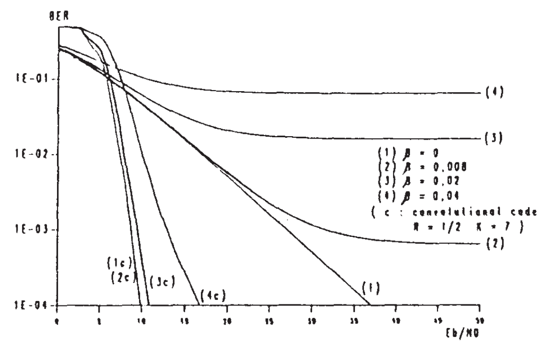

Figure 9

The performance of 4-PSK-COFDM in time-selective fading channels (differential demodulation)

The robustness of the COFDM signal as regards the problem of time coherence ($\beta = 0.04$ corresponds to a vehicle speed of 600 k.p.h at 900 MHz for a symbol duration of 80 μs), can be understood in the following manner. Considering the polar representation of a received pure frequency f_o, as a function of the time (fig.10a), it is clear that the derivative of the phase of the received signal with respect to time (also called random FM), will highly tend to increase when the amplitude of the signal is closed to zero. This strong correlation between the modulus of the transfer function of the channel $H(f_o,t)$ and the random FM clearly appears in figure 10b and 10c, and it can be shown [7] that the value of the random FM is given by :

$$F_m(t) = \frac{1}{2\pi}\frac{d\varphi(t)}{dt} = \frac{1}{2\pi}\frac{ReH.ImH' - ImH.ReH'}{\| H \|^2}$$

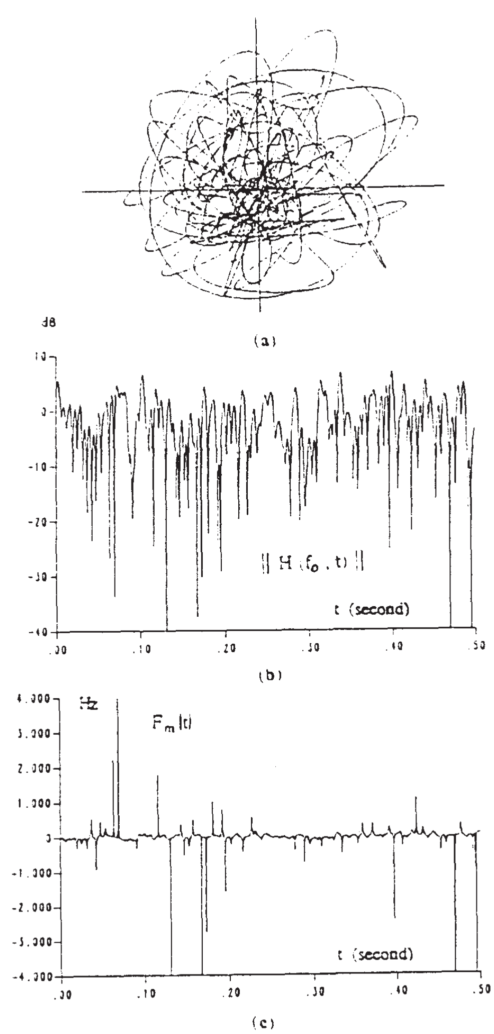

Figure 10

Channel characteristics variation in the time domain at a given frequency

a : Amplitude and phase variation at the given frequency f_o (Polar representation)

b : Amplitude as a function of the time $f_o = 900$ MHz . V = 120 k.p.h

c : Random FM as a function of the time same conditions as in b

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From this expression, one can see that, as the total energy appears in the denominator, large fluctuations of $F_m(t)$ will occur near deep fadings.

As it has been explained in section 4, the weightings of the Viterbi decoder are the real and the imaginary parts of $Y_{j,k}.Y^*_{j-1,k}/(1-i)\,\sigma_{j,k}^2$. This means that the soft decision channel decoder takes into account the effect of the attenuation at each time-frequency point ($\rho_{j,k}^2$), in such a way that the carriers affected by deep fadings, that are supposed to suffer important phase shifts, do not participate in a large extend to the maximum likelihood decoding process.

6.3 COFDM parameters choice

The conditions $\mu << 1$ and $\beta << 1$ imply that $f_{max}\ T_m << 1$. The spread factor defined as $f_{max}\ T_m$, or $[(\Delta t)_c\ (\Delta f)_c\ ]^{-1}$, depends only on the channel, and the COFDM system in the described version is effective for channels with a spread factor much less than one.

In fact, the parameters $1/t_s$ and T_s of the COFDM system must be chosen to verify approximatively the two conditions (considering $\Delta = t_s/4$) :

$$\mu < 0.1 \quad \left(\text{or } 1/t_s < \frac{1}{10T_m}\right)$$

$$\beta < 0.02 \quad \left(\text{or } T_s < \frac{1}{50f_{max}}\right)$$

Then, the spread factor of the channel must be less than $1.6\ 10^{-3}$, and leads to define the conditions of application of the adaptable CODFM system (fig. 11).

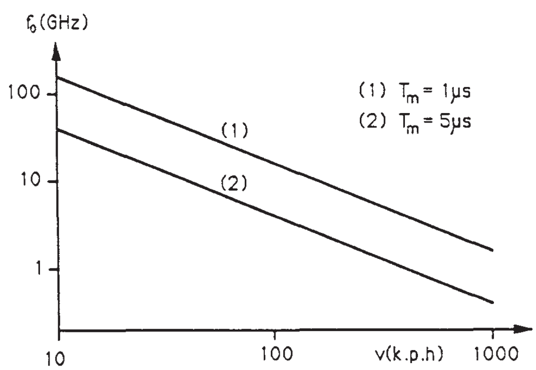

Figure 11

Maximum carrier frequency as a function of the velocity of the receiver

As an example, let us suppose a delay spread of 1 µs (typical case for suburban and urban areas) and a maximum Doppler shift of 100 Hz (which can be obtained with f_0 = 900 MHz and v = 120 k.p.h.). In this case, the spread factor is equal to 10^{-4} and the conditions become $12.5\ \mu s < T_s < 200\ \mu s$.

7. A DIGITAL SOUND BROADCASTING NETWORK

The main difficulties that arise during the introduction of a new broadcasting service are caused by the lack of frequencies in the radio-electrical spectrum. This spectrum belongs jointly to all potential users and must be used to the full. This is why, in addition to efficiency in terms of power and spectrum usage, flexible frequency planning must be taken into account.

Our research covers both terrestrial and satellite broadcasting and at present, we are looking into the following configurations :

- a local urban broadcasting service in the UHF band using 4 to 7 MHz bandwidth, that can be shared with television signals. The COFDM system requires very low power and it is therefore possible to re-use TV channels operating in adjacent areas.
- a regional or national broadcasting service operating in the 60-200 MHz band based on the implementation of a single-frequency network using basic 4 MHz bands. Since the COFDM system is designed to function in the presence of multipaths, it is possible to generate « active echoes » from a network of transmitters spread throughout a given territory. They would be temporally synchronized and would all transmit the same signal. The spread of the « equivalent transmission channel » impulse response is, of course, dependent on the physical distance separating the various transmitters. It is therefore necessary to use very long COFDM symbols (≈1 ms), with a guard interval capable of absorbing echoes from over 70 kilometers away. Increasing the size of the symbols clearly results in an increased number of carriers transmitted in a given bandwidth. This apparent constraint in fact corresponds to a slight increase in receiver complexity.
- a national satellite broadcasting service operating on the 0.5 - 2 GHz frequency band based on the utilization of basic 4 MHz bands. This possibility, which requires the exclusive allocation of a frequency band nevertheless includes all the well-known advantages of satellite broadcasting [13].

8. SYSTEM DEVELOPMENT

8.1 Choice of the modulation and coding parameters

The first real system was developed with the aim of validating the principles of the COFDM system in the case

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of UHF transmission [14]. It provides 16 stereophonic programmes, each with a rate of 336 kbit/s, in a total bandwidth of 7 MHz.

The various programmes are multiplexed in such a way as to minimize receiver complexity. As the receiver has to process only one programme at a time, the coding and modulation procedures have to be applied separately to the various sources. The multiplexing of the different programmes can be done in the frequency or in the time domain. In the case of frequency multiplexing, programme selection and demodulation can be performed jointly by decimating the FFT. Time multiplexing, which is more simple to implement, was retained for this first system.

The basic symbol has a duration of 80 µs and includes a guard interval of 16 µs. Each symbol constitutes a multiplex of 448 carriers separated by 15625 Hz, modulated separately in 4-PSK. The guard interval absorbs the multipaths in most situations. Moreover, the modulation symbols are sufficiently short to provide temporal channel coherence as regards the incoming signal, even at 200 kph and a carrier frequency of 1 GHz. This is a vital condition for demodulator operation, whether differential or coherent.

The signal is built up around a 24 ms frame structure corresponding to the juxtaposition of 300 symbols. The first symbol, which is free of modulation, is used to synchronize the receiver. The second symbol is a sine-sweep signal used as a phase reference for differential demodulation. The third symbol carries static data. The remaining 297 symbols are divided into 33 channels of 9 consecutives symbols. A channel corresponds either to a monophonic sound or to a data channel.

The channel coding system is a convolutional code, as described above ($R = 1/2$, $K = 7$, $d_{free} = 10$). Each channel has a transmission capacity of 168 kbit/s x 24 ms/frame, i.e. 4032 bit/frame. These 4032 bits are code-linked to form a block of 8064 bits. At the output of the convolutional encoder, the data are temporally interleaved over 16 frames, i.e. 384 ms ; frequency interleaving spreads the data over the 448 carriers.

8.2 Description of the receiver

The RF stage of the receiver is quite conventional. Channel filtering is performed in the intermediate frequency using a SAW filter with a 7.5 MHz band. The signal is then demodulated over two quadrature components at the central frequency $f_0 + 1/2T$ and the resulting I and Q signals are low-pass filtered at 4 MHz cut-off frequency and sampled at the rate of 8 MHz.

Synchronization is performed by filtering the envelope of the incoming signal, matched to the absence of modulation during the 80 µs of the first symbol of the frame. This first estimate provides a mean of prepositioning the FFT window. This positioning is then refined by a channel impulse response estimation, based on the matched filtering of the sine-sweep symbol.

This incoming signal is then processed using a signal processor which carries out a 512 complex points FFT on the symbols belonging to the selected programme. Differential demodulation based on the use of complex multiplication is applied to the 448 carriers of the multiplex. Viterbi decoding is performed by an ASIC especially developed by our institute.

The theoretical results obtained in the Gaussian and selective Rayleigh channels were confirmed by experimental measurements achieved with a multipath simulator. The implementation margin is equal to approximately one decibel.

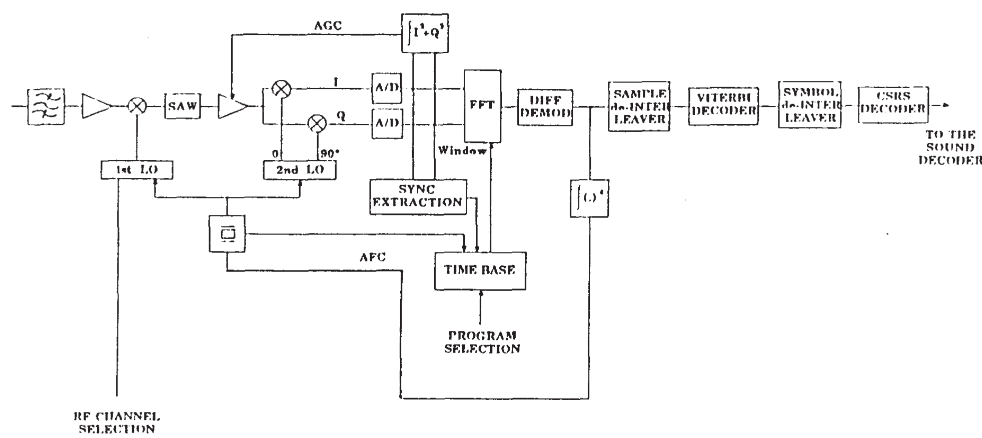

Figure 12

Synoptic diagram of the receiver

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CONCLUSION

Considering its spectrum efficiency, robustness and flexibility, the COFDM system can be regarded as an excellent candidate for an industrial development in the near future. In the context of the mass production domestic market, the feasibility of integrated circuits is now studied, and the first estimations are very promising.

We are also working on the possibility to use Ungerboëck type trellis coded modulations, associated with the orthogonal frequency division multiplexing technique, in order to increase the spectral efficiency of the system.

The first implementation of the 4-PSK-COFDM system has been demonstrated at the WARC-ORB 88 conference in Geneva [15], associated to an audio bit-rate reduction system designed by the Institut für Rundfunktechnik (FRG), and based on a sub-band coding technique. Since that time, the transmission system has been working for demonstration in the town of Rennes (France), using a studio compatible sub-band audio coding process [16] developed by our institute, which compands the data rate of a monophonic sound down to 128 kbit/s. In the present state of development, the spectral efficiency of the system is thus equal to about 300 kHz per stereophonic sound.

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