SPECIFICATION AND DESIGN OF A PROTOTYPE FILTER FOR FILTER BANK BASED MULTICARRIER TRANSMISSION

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ABSTRACT

The specifications of filter banks for multicarrier transmission systems with a large number of subchannels are discussed, with application to xDSL and power line communication in mind.The near PR modulated approach is considered and the importance, for the system, of the protype filter delay is stressed. An existing design technique known to be particularly relevant to the context is revisited from a frequency sampling perspective. The performance results in terms of subchannel noise floor and delay are given for several filter lengths and an experimental validation is provided. Finally, an improvement to the design technique is proposed, which brings a gain of 3.3 dB in subchannel interference power level.

1. INTRODUCTION

A filter bank based multicarrier transmission system employs two filter banks, a synthesis filter bank (SFB) at the transmitter side and an analysis filter bank (AFB) at the receiver side [1]. In such a system, there is no need for guard times to separate consecutive symbols, as in the Digital MultiTone (DMT) technique implemented in Asynchronous Digital Subscriber Loop (ADSL) equipments for example. The consequence is an increase in efficiency and, also, a significantly enhanced robustness. Since the subchannels are well separated, high level narrow-band disturbing signals or jammers affect only a few subchannels , efficient subchannel equalisers can be used, constraints on synchronisation are relaxed and the initialisation sequence at the beginning of a connection can be shortened.

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Filter banks can be designed with the perfect reconstruction (PR) or near PR property [2]. PR filter banks are appropriate for subband coding, for example.

However, here, as shown in figure 1, an analog channel is present between synthesis and analysis filters, which introduces distortion and noise. In such a context, PR would imply perfect equalisation in the receiver, which is unrealistic. Therefore, there is no need to introduce extra design constraints and modulated near PR filter banks are considered.

The paper is organised as follows. In section 2, the specifications of the filter banks are reviewed and justified. In section 3, it is shown that the technique proposed in [3] can be viewed as a frequency sampling method, which leads to the improvement presented in section 4. The conclusion discusses the trade-off between system delay and noise floor, based on the performance results.

2. SPECIFICATION OF THE PROTOTYPE FILTER

The transmission channel is divided into N subchannels, with frequency spacing 1/2N, assuming the sampling frequency is fs=1. In each subchannel, the symbols are transmitted at a rate equal to twice the subchannel spacing, namely 1/N. If OQAM (Orthogonal Quadrature Amplitude Modulation) modulation is used, the input is a complex number and the symbols are transmitted alternatively on the real part and the imaginary part [4]. For a DSL type of application, the sampling frequency can be 2048 kHz and the number of subchannels N=256. With this number of subchannels, the modulated filter bank approach is convenient. It is based on a low-pass prototype filter whose transition band is centered on the frequency fc=1/4N.

The prototype filter controls the phase and amplitude distortions in the subchannels and the interference between subchannels. Phase distortion is eliminated if the prototype filter is linear phase. In principle, the square of the frequency response must satisfy the Nyquist criterion for data transmission and amplitude distortion produces intersymbol interference in the receiver. However, if a subchannel equaliser is employed, some flexibility can be introduced.

A key specification in digital transmission is the delay. An overall delay budget is alloted to the system and it is shared by the various functions in the transmitter and receiver. The main contributions come from modulation, optimal detection and error correction. For example, in single carrier transmission like QAM and CAP (Carrier-less Amplitude Phase), most of the delay budget is used for maximum likelyhood decoding and error correction. In contrast, multicarrier techniques, particularly when filter bank based,consume a significant portion of the delay budget for modulation and demodulation. Therefore, there is a strong pressure to minimise the delay of the prototype filter and the selection of the number of coefficients L is a trade-off between delay and filter performance, mainly stop-band attenuation.

A minimum phase FIR filter can be envisaged, but the ratio of stop-band to pass-band is very high in that case, because the middle of the transition band is fc=1/4N, N is large and almost all the filter zeros are in the stop-band and on the unit circle. Thus, there is little to be gained by eliminating the zeros out of the unit circle.

One way to reduce the filter length L and the delay is to expand the transition band. With the filter bank approach, considering subchannel i, the orthogonality principle is kept for the neighbouring subchannels i+1 and i-1. Therefore, interference comes from subchannels i±2. Consequently, the maximum transition bandwidth is 1/2N, which is also the subchannel spacing.

In the system, two neighbouring subchannels overlap and the corresponding interference function is

$$I_1(f) = H(f)\, G\left(\frac{1}{2N} - f\right) \qquad (1)$$

with H(f) and G(f) being the protype filter responses for transmitter and receiver respectively. Interference is cancelled at the symbol sampling times if $I_1(f)$ is symmetrical about the frequency fc=1/4N, which imposes G(f)=H(f), and if the filter delay is an integer multiple of the sampling period 1/fs, which imposes that the length L be an odd number [4].

Then, the delay of the transmitter-receiver cascade is

$$D = 2\,\frac{L-1}{2}\,\frac{1}{fs} = (L-1)\frac{1}{fs} \qquad (2)$$

and the filter length is determined by the desired level of interference between the subchannels.

As mentioned earlier interference occurs between subchannels whose difference of indices is a positive even number. It is produced at both transmitter side and receiver side.The situation at the transmitter is depicted in figure 2. Since the emitted signal is the sum of the outputs of all the subchannel filters, in subchannel i, the residuals of subchannel signals with indices i±2k (1≤k≤N/2) add up.

Assuming the subchannels are fed with independent unit power data symbols, the corresponding interference power is estimated by

$$P_{ie}=2\int_{-1/2N}^{1/2N}\left|H(f)\right|^2\left(\sum_{k=1}^{N/2}\left|H\left(f-\frac{2k}{N}\right)\right|^2\right)df \qquad (3)$$

At the receiving side, a similar situation occurs, because the filter bank imperfectly selects channel i and the residuals of subchannels with indices i±2k (1≤k≤N/2) are aliased by the subsampling process. An interference power equal to P_{ie} results. As concerns the addition of the interference signals from emission and reception, although they concern the same signals, one can consider that, due to the presence of the transmission channel transfer function introduced in-between, they add up in power. Therefore, the noise floor in each subchannel can be taken as

$$N_0 = 2\,P_{ie} \qquad (4)$$

For example, for ADSL, some operators consider that the noise floor should not exceed – 45dB, which, through relation (3), defines the attenuation of the prototype filter. It must be emphasized that this noise floor is the price which has to be paid for the absence of global orthogonality between the subchannels.

3. DESIGN OF PROTOTYPE FILTER

The prototype filter can be designed with the help of classical optimisation techniques [5]. However, considering that the length L can reach several thousands, a direct technique seems appropriate, at least as a starting point, and the method proposed in [3] has been shown to be particularly relevant to the problem. In fact, it will be shown now that the technique can be interpreted as a frequency domain technique.

Let us consider an integer K and a set of KN samples H_k $(0 \le k \le KN-1)$ in the frequency domain, such that

$$H_0 = 1 \qquad (5)$$

$$H_k^2 + H_{K-k}^2 = 1 \quad ; \quad H_{KN-k} = H_k \quad ; \quad 1 \le k \le K-1$$

$$H_k = 0 \qquad ; \qquad K \le k \le KN-K$$

Generally, the number of channels N is even. The Nyquist criterion is satisfied by these frequency samples and the corresponding filter coefficients h_i ($0 \le i \le KN-1$) are obtained by inverse DFT (Discrete Fourier Transform). Since the number of coefficients L must be an odd number, the coefficient $h_{NK/2}$ can be cancelled by imposing

$$H_0 + 2\sum_{k=1}^{K-1}(-1)^k H_k = 0 \qquad (6)$$

For K=3 and K=4, equations (5) and (6) define a determined system and the frequency samples are

$H_1 = 0.911438 \quad ; \quad H_2 = 0.411438$

and

$H_1 = 0.971960 \quad ; \quad H_2 = \sqrt{2} \quad ; \quad H_3 = 0.235147$

Those are the results obtained in [3], which gives an iterative procedure that can be used for K≥5. Alternatively, a systematic search can be carried out, starting, for example, from the ideal frequency response samples

$$H_k = \cos \pi \frac{k}{2K} \qquad (1 \le k \le K) \qquad (7)$$

to minimise the interference power P_{ie} in (3).

Table 1 gives the interference power computed for several values of parameter K, as well as the corresponding system delay.

L	4N-1	6N-1	8N-1	10N-1
Interf. Power (dB)	–17.8	-37.0	-50.1	-73.6
Delay (x 1/fs)	4N-2	6N-2	8N-2	10N-2

Table 1. Interference power versus filter length

For the length L=8N-1 and N=512, the coefficient values have been introduced in an implementation of the transmission system. The noise floor has been measured and found to be –49dB. This is in good agreement with the calculated value –50.1 dB and validates the interference power calculation.

4. IMPROVED FILTER DESIGN

Another option than having 2KN-1 coefficients and using equation (6) can be taken. In fact, the filter can have 2KN+1 coefficients, with

$$h_{NK/2} = h_{-NK/2} = ½[H_0 + 2\sum_{k=1}^{K-1}(-1)^k H_k] \qquad (8)$$

A systematic search can, then, be performed, to take advantage of the removed constraint. The filter frequency response obtained that way for K=4 and L=8N+1 coefficients, with N=64, is given in figure 3. The interference power obtained is –53.4 dB.

Note that, with the frequency sampling approach, the filter transfer function has a double zero at frequency 0.5 [5]. Otherwise, the distribution of the zeros is uniform over the stop-band and the envelop of the frequency response decreases monotonically. This is a desirable feature to better protect each subchannel from distant high level jammers.

5. CONCLUSION

For the designer of a multicarrier transmission system, the results given in Table 1 are crucial, because they provide the correspondence between the system delay due to the filter banks and the noise floor in each subchannel which limits the capacity of the system.

For example, in the context of ADSL, the length L=8N+1 is satisfactory if the noise floor is set to –45dB. Assuming the sampling frequency is fs=2048 kHz and N=256, the delay is d=1 ms, which is compatible with the one way transfer delay of 2 ms specified in recommendation ITU-T / G.992.1 [6].

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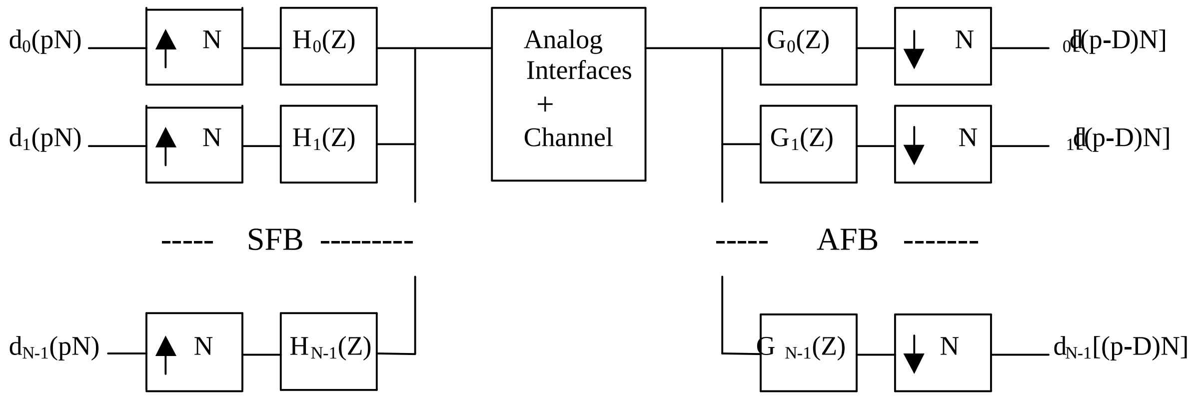

Fig.1 Principle of filter bank based multicarrier transmission

Fig.2. Interference between subchannels

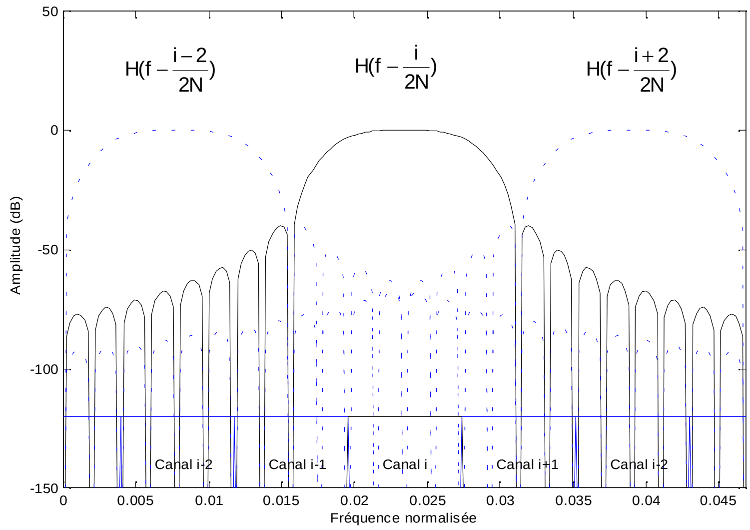

Fig.3. Frequency response of length L=8N+1 prototype filter