行政院國家科學委員會補助專題研究計畫成果報告

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計畫類別：☑個別型計畫　　□整合型計畫
計畫編號：NSC89－2213－E－009－133
執行期間：88 年 8 月 1 日至 89 年 7 月 31 日

計畫主持人：張文鐘教授
共同主持人：

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執行單位：國立交通大學電信工程學系

中　華　民　國　89　年　9　月　1　日

transform of a block may depend not only on that block but also on its neighboring blocks.

In the DMT system, the modulation is a QAM. However, designing a DWMT waveform for a QAM modulation will face more constraints than for an offset QAM modulation [2]. Therefore, we discuss only the OQAM case. The M -band OQAM-OFDM system can be seen as applying M OQAM to a set of adjacent frequencies. The carrier frequencies are separated by $\frac{1}{T}$, where T is the symbol period. This is to say that the bandwidth of each tone is $\frac{1}{T}$. So, the effective baseband bandwidth is 1/2(1/T). This is the Nyquist bandwidth.

Denote $c_{k,n}$ as the complex symbol transmitted in the n th band at time $t = kT$. The real and imaginary parts of $c_{k,n}$ are separated in time by half a symbol interval, $\frac{T}{2}$. Denote $\hat{c}_{k,n}$ as the corresponding received symbol and $h(t), g(t)$ the impulse responses of the transmitting and receiving filters respectively. Note that they are real functions and FIR. Also to be noted is that the carriers have a phase difference of $\frac{\pi}{2}$ between adjacent tones. The point is how to design the transmitting and receiving filters such that the individual symbols in the receivers will not interfere with each others.

三、研究方法及成果

To derive the conditions for zero inter-channel and inter-symbol interference, consider the signal path from channel $m'+n$ to channel n of the OQAM-OFDM system. For the orthogonal condition, we represent the relations by the four following equations.

$$\left[\mathrm{Re}\left\{h(t-kT)e^{j(\frac{2\pi}{T}t+\frac{\pi}{2})m'} * g(t)\right\}\right]_{t=0} = \delta(m',k)$$

$$\left[\mathrm{Re}\left\{jh(t-kT+\frac{T}{2})e^{j(\frac{2\pi}{T}t+\frac{\pi}{2})m'} * g(t)\right\}\right]_{t=0} =$$

$$\left[\mathrm{Im}\left\{h(t-kT)e^{j(\frac{2\pi}{T}t+\frac{\pi}{2})m'} * g(t-\frac{T}{2})\right\}\right]_{t=0} = 0$$

$$\left[\mathrm{Im}\left\{jh(t-kT+\frac{T}{2})e^{j(\frac{2\pi}{T}t+\frac{\pi}{2})m'} * g(t-\frac{T}{2})\right\}\right]_{t=0} = \delta(m',k)$$

where $m' = 0,1,2,\ldots,M$. For simplicity, we choose $g(t)$ as the time reverse version of $h(t)$, the matched filter, and restricting $h(t)$ to be real and symmetric. Then for the above equations, only the following one is needed to be concerned with $m' = 2m$.

$$G_{k,m} = \int_{-\infty}^{\infty} h(t-kT)h(t)\cos(\frac{2\pi}{T}2mt)dt = \delta(m',k) \quad (1)$$

where $m' = 2m$ and $m = 0,1,2,\ldots,\left[\frac{M}{2}\right]$.

With the above orthogonality requirement, the goal of the design is to concentrate the energy of the waveform to derive a good low-pass filter. Assuming J being the energy of $h(t)$ within the bandwidth $\left[-\frac{2\pi}{T},\frac{2\pi}{T}\right]$. In time domain, it extends within (-L,L). That is

$$J = \int_{-\frac{2\pi}{T}}^{\frac{2\pi}{T}} |H(\omega)|^2 d\omega$$

or

$$J = \int_{-L}^{L}\int_{-L}^{L} h(t)h(s)\frac{\sin\frac{2\pi}{T}(t-s)}{\pi(t-s)}dtds$$

Now our goal is to maximize J under the constraints of Eq(1). To solve this problem, we first discretize the waveform h(t) such that J is decided by $h(t_d)$, $t_d \in [-L,L]$. We can regard J is a function with variables $h(t_d)$. Then we form the following equation

$$J'(h(t_d)) = J(h(t_d)) + \sum_{m',k}\alpha_{m',k}\left[\int_{-\infty}^{\infty} h(t-kT)h(t)\cos(\frac{2\pi}{T}2mt)dt - \delta(2m,k)\right]$$

To maximize the above equation, we derive

$$\frac{\partial J(h(t_d))}{\partial h(t_d)} = \frac{\partial}{\partial h(t_d)}\left[\int |H(\omega)|^2 d\omega\right]$$

$$= 2\int h(t)\cdot\frac{\sin\omega(t_d-t)}{t_d-t}dt$$

$$\frac{\partial}{\partial h(t_d)}\left[\int h(t-kT)h(t)\cos\left(\frac{4\pi m}{T}t\right)dt-\delta(2m,k)\right]$$

$$=\left[h(t_d+kT)+h(t_d-kT)\right]\cdot\cos\left(\frac{4\pi m}{T}t_d\right)$$

Therefore, the constraint equation differentiated by $h(t_d)$ is equals to

$$\frac{\partial J'(h(t_d))}{\partial h(t_d)}=2\int h(t)\cdot\frac{\sin\frac{2\pi}{T}(t_d-t)}{(t_d-t)}dt+$$

$$\sum_{m,k}\alpha_{m,k}\left[h(t+k_d)+h(t-t_d)\right]\cdot\cos\left(\frac{4\pi m}{T}t_d\right)=0$$

That is

$$2\int h(t)\cdot\frac{\sin\frac{2\pi}{T}(t_d-t)}{\pi(t_d-t)}dt=$$

$$\sum_{m,k}\lambda_{m,k}\left[h(t+k_d)+h(t-t_d)\right]\cdot\cos\left(\frac{4\pi m}{T}t_d\right)$$

where $\lambda_{m,k}=-\frac{\alpha_{m,k}}{\pi}$ $\quad t_d:[-L,L]$

四, 結果與討論

The waveform is set to be a linear combination of a set of cibic splines. We compress a cubic spline whose length is 4 into $[0,X]$ and sample it with nine points including the two ends. The interval between the adjacent samples is $\frac{X}{8}$. A cubic spline is defined as

$$\Psi_0(t)=\begin{cases}\frac{t^3}{6} & ,0\le t<1\\ \frac{-3t^3+12t^2-12t+4}{6} & ,1\le t<2\\ \frac{3t^3-24t^2+60t-44}{6} & ,2\le t<3\\ \frac{(4-t)^3}{6} & ,3\le t<4\\ 0 & ,otherwis\end{cases}$$

and $\Psi_i(t)=\Psi_0(t-i)$

Assuming the corresponding weighting for Φ_i is a_i, we then represent $h(k)$ as $h(k)=\sum a_i\Phi_i(k)$ where $k=\frac{X}{8}d+\frac{X}{16}$, $d\in Z$.

The result with two spans is shown in Fig.1 with M=8,16,24 and filter length 32,64,96. The result with three spans is shown in Fig.2 with M=9,15,21 and filter length 54,90,126. As can be seen that the frequency responses of these filters are all better than that of the rectangular filter and also the side lobes are almost 40db below the main lobe.

We have tested these waveforms in a simulated DMT system with perfect channel and the result is as expected that no ICI and ISI occurs. For a no ideal channel, the result of the ICI and ISI is seen to be less than the case with the rectangular waveform. The cost of such improvement is the increase of the modulation complexity associated with a LOT or ELT transform. When the time comes where channel bandwidth efficiency is more valuable than the computing power, this scheme will prove useful.

五、參考文獻

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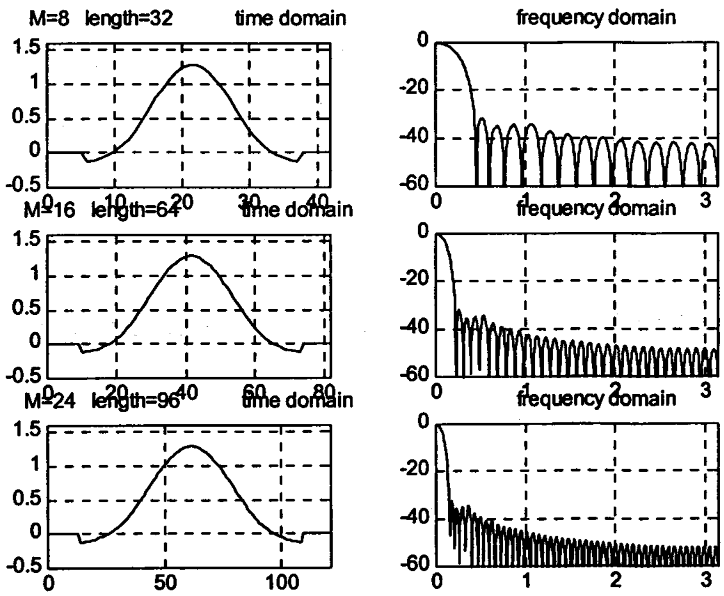

Fig. 1 Filter with 2 spans

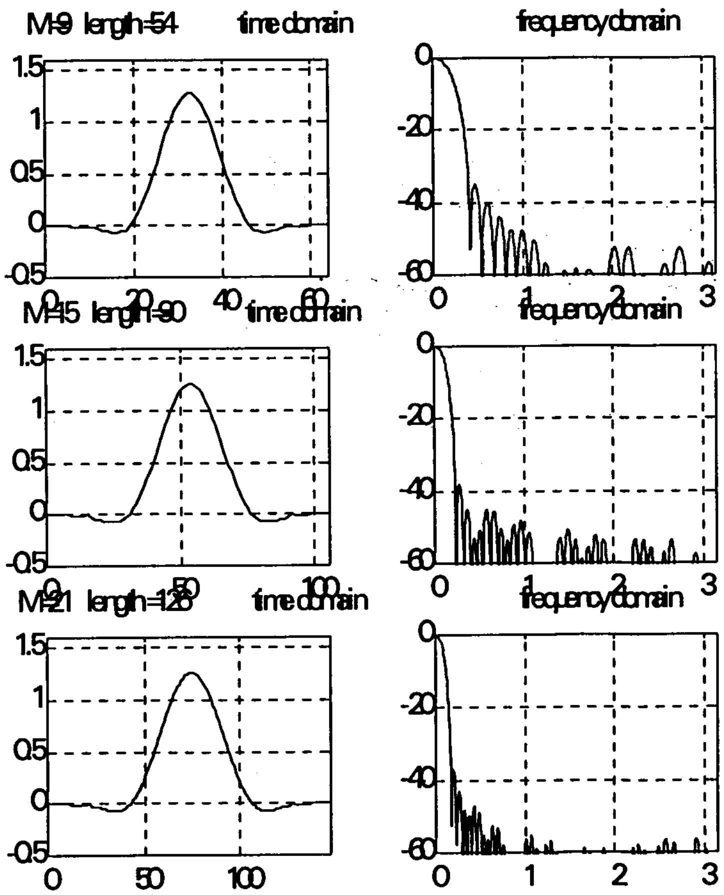

Fig. 2 Filter with 3 spans