

CO-CHANNEL INTERFERENCE CANCELLATION FOR HDTV RECEIVERS

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ABSTRACT

A new method of co-channel interference rejection for digital television receivers is presented that uses a different rejection filter from the comb filter that was used in the prototype built by Zenith. This filter is optimized for rejection of co-channel NTSC interference *in the presence of white noise* and hence suffers a penalty of only 0.4 dB in AWGN as compared to 3.5 dB with the comb. The receiver structure with this filter, including required equalizer and trellis decoder modifications is presented, along with simulation results showing the improvement in performance in co-channel-plus-AWGN interference and hence in coverage area.

1. INTRODUCTION

The 8-VSB trellis-coded system [1] that has been adopted as the standard for digital terrestrial television transmission in the US has to initially co-exist with analog NTSC transmissions. Digital transmissions will be on presently unused “taboo” channels that will suffer from co-channel NTSC interference in certain locations which are midway between an analog and a digital transmitter transmitting on the same frequency. Hence, in order to increase the coverage area of the digital transmission, digital television receivers need to have some means of co-channel interference cancellation. The 8-VSB standard as adopted does not have any means in the transmitter, other than a symbol interleaver, to assist in co-channel cancellation at the receiver.

The prototype receiver built by Zenith had a 12 symbol-delay “comb filter” in the receiver for co-channel cancellation. This filter worked well in the presence of NTSC alone but suffered a 3.5 dB loss when AWGN was present in addition to moderate levels of NTSC interference. This led to a system where the receiver detected the level of NTSC and switched the comb filter in when the level was high. As described in [2], this posed problems in the field where accurate detection of NTSC interference levels is difficult in the presence of other impairments like impulse noise, multipath etc. It was observed in field tests that the comb filter was switched in erroneously in situations where there was not a high enough level of NTSC interference, with the resulting loss of 3.5 dB in threshold SNR.

In this paper, we propose a different kind of NTSC rejection filter which is better suited for NTSC cancellation

in the presence of AWGN and suffers only 0.4 dB loss in performance in AWGN-only. This filter was first proposed in [3] and described there in a precoder-filter arrangement, i.e., the filter coefficients were used in the transmitter to precode the signal prior to transmission using Tomlinson-Harashima (TH) precoding. While this approach gives the best performance in co-channel, it requires a modification in the transmission standard, which is not possible anymore. Hence, in this paper we develop a scheme whereby a similar filter can be used only in the receiver, in place of the comb filter. Section 2 briefly describes the 8-VSB receiver architecture with the comb filter. Section 3 describes the new scheme with the rejection filter and modifications to the equalizer and trellis decoder. Simulation results of both the cancellation schemes are presented and discussed in Section 4. Finally, conclusions are presented in section 5.

2. 8-VSB WITH COMB FILTER

The detailed specification of the US standard for terrestrial transmission using 8-VSB can be found in [1] and [4]. Incoming data packets are scrambled, Reed-Solomon (RS) encoded, convolutionally interleaved on a byte basis, trellis encoded and symbol interleaved before transmission. The symbol interleaving and trellis encoding employed in the transmitter is done in such a way that there are 12 identical parallel trellis encoders and every 12th symbol in the transmitted data stream comes from the same encoder. When the received data stream is then passed through a 12-symbol-delay comb filter in the receiver, the effective delay seen by each trellis decoder is only one symbol. This facilitates the trellis decoding at the receiver. The received sequence r_k is, in the most general case, distorted by multipath, AWGN and co-channel NTSC and can be expressed as:

$$r_k = \sum_{i=0}^{L_h-1} h_i a_{k-i} + w_k + n_k \quad (1)$$

where a_k is the transmitted 8-VSB sequence (symbols from the set $(\pm 1, \pm 3, \pm 5, \pm 7)$), h_i is the multipath of length $L_h - 1$, w_k is the AWGN with variance σ_w^2 and n_k is the co-channel NTSC interference with a correlation $R_n(p) = \mathbf{E}[n_k n_{k-p}]$. The output of the comb filter, x_k , can be expressed as:

$$x_k = r_k - r_{k-12} = \sum_{i=0}^{L_h-1} h_i a'_{k-i} + w'_k + n'_k \quad (2)$$

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where $a'_k = a_k - a_{k-12}$ is distributed over the 15 levels $(0, \pm 2, \pm 4, \pm 6, \pm 8, \pm 10, \pm 12, \pm 14)$, the co-channel component n'_k has a lower variance than the received co-channel n_k but the gaussian noise component w'_k has a 3 dB higher variance than the received AWGN w_k . This 3 dB increase in the noise variance, coupled with the fact that the noise is no longer white, causes a net loss of 3.5 dB in AWGN performance at the trellis decoder output. Hence, the comb filter cannot be used in situations where the SNR is low. Following the comb filter is a decision feedback equalizer (DFE) whose input is x_k . Since the signal component of x_k is 15-level as shown above, the slicer in the DFE now operates on these 15 levels to ensure that the equalizer does not equalize the comb filter response. The trellis decoder following the DFE uses an expanded trellis to compensate for the intersymbol interference (ISI) introduced into the transmitted symbol stream by the comb filter.

Figure 1 shows the spectrum of a typical NTSC signal, the color-bars signal, and the frequency response of the comb filter. The spectrum of an NTSC signal does not change appreciably with content, i.e., the video signal for a regular picture will have major spectral components that are similar to those of the static color-bars signal. The comb has nulls at frequencies where the NTSC signal has maximal energy, viz., the picture, sound and color carriers. However, it also has additional nulls at frequencies where there is little-to-no NTSC energy to be attenuated. This creates additional ISI and noise enhancement.

3. CO-CHANNEL REJECTION FILTER

In [3], a method of constrained optimization using Lagrange multipliers was described for designing a filter that could be used for co-channel NTSC cancellation without excessive noise enhancement. In this section we will describe how such a filter can be used in the receiver alone for co-channel NTSC cancellation.

Figure 2 shows the receiver architecture with the co-channel rejection filter. The filter is designed to minimize the NTSC variance at its output while limiting the noise enhancement to about 0.4 dB. The first coefficient is 1 and the other coefficients are much smaller. The coefficients for a 37-tap filter designed to meet these specifications are shown in Table 1. These coefficients have been quantized to 5 bits in order to reduce the complexity. After quantizing, there are only 25 non-zero coefficients even though the filter length is 37. This filter can be implemented with only adders (instead of multipliers) using the canonical signed-digit (CSD) representation.

Figure 3 shows the frequency response of the filter of Table 1 relative to the NTSC spectrum. Compared to Figure 1 we see that the rejection filter has nulls only where they are required (at the picture, color and sound carrier frequencies), unlike the comb filter. Hence, there is less distortion of the signal when NTSC co-channel is absent. This allows the rejection filter to be left in all the time with a loss of only 0.4 dB in SNR, compared to the comb which suffers a 3.5 dB penalty and therefore needs to be switched out in the absence of co-channel NTSC.

Referring back to Figure 2, the received signal r_k is first input to the rejection filter g_k which is an FIR filter

g_0	1.00000	g_{13}	0.00000	g_{26}	-0.06250
g_1	0.00000	g_{14}	-0.06250	g_{27}	0.00000
g_2	-0.06250	g_{15}	0.00000	g_{28}	0.00000
g_3	0.03125	g_{16}	0.00000	g_{29}	0.03125
g_4	0.00000	g_{17}	0.06250	g_{30}	0.03125
g_5	0.06250	g_{18}	0.03125	g_{31}	0.06250
g_6	0.06250	g_{19}	0.06250	g_{32}	0.03125
g_7	0.06250	g_{20}	0.00000	g_{33}	0.03125
g_8	0.00000	g_{21}	0.03125	g_{34}	-0.03125
g_9	0.03125	g_{22}	-0.06250	g_{35}	0.00000
g_{10}	-0.06250	g_{23}	0.00000	g_{36}	-0.06250
g_{11}	0.00000	g_{24}	-0.09375		
g_{12}	-0.09375	g_{25}	-0.03125		

Table 1: 5-bit coefficients for a 37- tap NTSC rejection filter.

with taps $[1, g_1, g_2, \dots, g_{L_g-1}]$. The output of the rejection filter, x_k is input to the DFE. Unlike a conventional DFE, in Figure 2 the DFE does not try to restore the data stream a_k because if it did it would have to undo the effect of the filter g_k . Instead, the DFE tries to reconstruct the sequence c_k which is defined as follows:

$$c_k = g_k \star a_k = a_k + \sum_{i=1}^{L_g-1} g_i a_{k-i} \quad (3)$$

where $\star$ denotes convolution. Hence, c_k is the response of the known filter to the data stream a_k . The input to the equalizer can now be written as:

$$x_k = c_k \star h_k + n_k \star g_k + w_k \star g_k \quad (4)$$

The filter g_k is designed as described in [3] so as to minimize the co-channel component at its output ($n_k \star g_k$) while enhancing the gaussian noise component ($w_k \star g_k$) by only 0.4 dB.

The output of the equalizer $\tilde{c}_k$ is defined as follows:

$$\tilde{c}_k = f_k \star x_k - b_k \star \hat{c}_k \quad (5)$$

where f_k are the forward equalizer taps, b_k are the feedback taps and $\hat{c}_k$ is the "sliced" version of $\tilde{c}_k$. During the training sequence, a_k is known and hence c_k can also be calculated since the filter coefficients g_k are fixed and known. The error signal $e_k = c_k - \tilde{c}_k$ is then used to drive the LMS algorithm, and c_k is the input to the feedback filter. Now, after the equalizer has converged on the training sequence and data is being transmitted, a_k is no longer known, and hence c_k is unknown too. However, the feedback filter still requires $\hat{c}_k$ as its input. Since c_k does not have discrete levels, a modified "slicer" is derived as follows. From equation (3):

$$a_k = c_k - \sum_{i=1}^{L_g-1} g_i a_{k-i} \quad (6)$$

Hence, in the modified slicer $\tilde{a}_k$ can be reconstructed from the equalizer output $\tilde{c}_k$ and past decisions $\hat{a}_k$ as follows:

$$\tilde{a}_k = \tilde{c}_k - \sum_{i=1}^{L_g-1} g_i \hat{a}_{k-i} \quad (7)$$

Now, $\tilde{a}_k$ can be sliced in the usual way with an 8-level slicer to give $\hat{a}_k$. Finally, the input to the feedback taps of the equalizer, $\hat{c}_k$ is obtained as follows:

$$\hat{c}_k = \hat{a}_k + \sum_{i=1}^{L_g-1} g_i \hat{a}_{k-i} \quad (8)$$

$\tilde{c}_k$ is the input to the next stage of the receiver, the trellis decoder. The modified slicer ensures that the DFE only equalizes the ISI introduced by the multipath channel h_i , and not that introduced by the rejection filter, in the same way that the 15-level slicer in the comb filter implementation ensures that the DFE does not equalize the ISI introduced by the comb.

The trellis decoder now has to remove the intersymbol interference (ISI) introduced into the symbol stream by the rejection filter, i.e., it has to recover the sequence a_k from its input sequence $\tilde{c}_k$, which can be expressed as follows:

$$\tilde{c}_k = a_k + \sum_{i=1}^{L_g-1} g_i a_{k-i} + N_k \quad (9)$$

where N_k is the residual noise after the equalizer which is composed of gaussian noise, NTSC interference and unequalized ISI. The maximum likelihood trellis decoder requires the use of an expanded trellis with a larger number of states to accommodate the memory introduced by the filter. A suboptimal, but computationally less intensive way of performing the trellis decoding is to instead implement the delayed-decision-feedback-sequence-estimation (DDFSE) method that was proposed in [5], where the following branch metric is used for each state j in the trellis at time k :

$$\left[\tilde{c}_k - a_k - \sum_{i=1}^{L_g-1} g_i \hat{a}_{k-i,j} \right]^2 \quad (10)$$

where the sequence $\hat{a}_{k-i,j}$, $i = 1, \dots, L_g - 1$ is the survivor symbol sequence associated with state j in the trellis. Thus, this scheme does not expand the number of states in the original trellis, but instead introduces decision-feedback in each of the trellis states.

4. SIMULATION RESULTS

Figure 4 shows the co-channel-NTSC-and-AWGN performance of 3 systems: (a) the equalizer-only system with a DFE having 64 forward and 192 feedback taps as described in [1], (b) the comb filter only system and (c) the proposed rejection filter system. The SNR is the signal-to-AWGN ratio and the DUR is the desired-to-undesired-ratio (i.e signal-to-co-channel-NTSC). The curve is plotted for an output byte error rate from the trellis decoder of about 1.4×10^{-2} . This corresponds to a packet error rate of 2×10^{-4} after the RS decoder, which has been determined to be the required threshold of visibility (TOV) for a HDTV system. The co-channel interference used in the simulation was a NTSC color-bars signal sampled at the 8-VSB symbol rate of 10.76 MHz. For each of the three systems, the region to the right of the curve corresponds to a byte error rate below TOV and the region to the left is above TOV. Hence,

the goal of any co-channel cancellation scheme should be to maximize the area to the right of its performance curve.

With the equalizer only, the system can tolerate co-channel NTSC at a DUR of about 12.7 dB when the SNR is 25 dB. The comb filter on the other hand allows the receiver to operate with a DUR of -2.7 dB at the same SNR. However, the equalizer-only system has a AWGN threshold of about 15 dB whereas the corresponding threshold for the comb is 18.5 dB due to the noise enhancement introduced by the comb. Hence, the Zenith prototype employed a hybrid system to take advantage of the performance of systems (a) and (b) in different SNR-DUR combinations and devised a receiver whereby the comb filter would switch in at point X in Figure 4. The performance of this hybrid system is marked by squares in Figure 4. The problem with this system is that in most real situations, significant co-channel can be present at the fringe areas of reception where the SNR is also low (15-20 dB), i.e., region (A) which is to the left of the hybrid system performance curve where the byte error rate is above TOV. Moreover, in the field, actual determination of co-channel conditions is very difficult and as described in [2], there were observed cases where the comb filter was switched in when it should not have been, and vice versa. It is clear from Figure 4 that in the neighborhood of the optimal switch point X, any error in the decision of comb switching will lead to the system moving from below TOV to above TOV. This is due to the steep slope of curve (b), the 3.5 dB loss in AWGN whenever the comb is switched in erroneously and the almost 12 dB loss in DUR when the comb is switched out when it should be in.

The rejection filter performance on the other hand is smoother across the entire SNR-DUR range, as shown by the curve (c) marked with circles in Figure 4. While the DUR for a SNR of 25 dB is 0.4 dB as compared to -2.7 dB of the comb filter, the performance in the 15-20 dB SNR range is clearly superior. At the other end, the SNR threshold for DUR of 35 dB is 15.4 dB, which is a loss of only 0.4 dB as compared to the 3.5 dB loss of the comb. Hence, the rejection filter can be employed all of the time without any of the problems inherent in comb filter switching. The increased coverage area obtained with the rejection filter can be explained as follows. Receivers in region (A) will be above TOV with the hybrid Zenith prototype system but below TOV with the rejection filter system. In regions (B) and (C), the Zenith prototype system will be below TOV whereas the rejection filter system will be above TOV. However, the *number* of receivers in region (A) is greater than that in regions (B) and (C). Moreover, receivers in region (B) can be reclaimed by the rejection filter system by switching the filter in at point Y instead of leaving it in all the time.

5. CONCLUSIONS

The comb filter is optimized for a high SNR-low DUR environment which does not exist in a real application since it would require the digital receiver to be very close to both the digital and analog transmitters. The majority of digital receivers that will be subject to co-channel interference will be on the fringe areas of reception of the digital signal where the SNR is low. In this region the comb filter

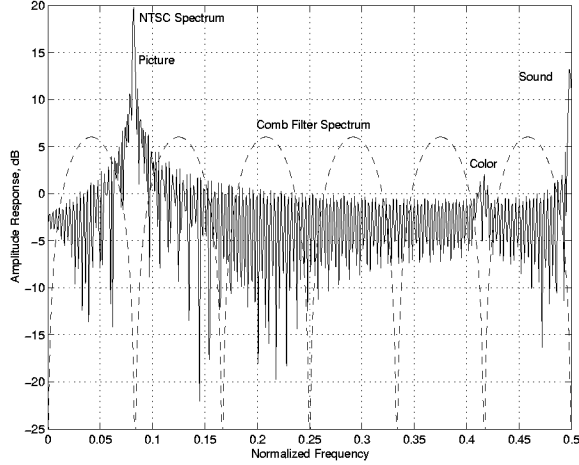


Figure 1: Spectra of co-channel NTSC and comb filter

does not perform well. In this paper, an alternative to the comb filter approach was presented that was shown to work better over a wider SNR-DUR range than the comb filter system and hence enhances the coverage area of the digital signal. The complexity of this new receiver structure is very comparable to the comb filter with enhanced trellis since the use of the DDFSE keeps the trellis decoder complexity essentially unchanged. The use of CSD coefficients for the filter further reduces the complexity since the filter can be implemented with adders instead of multipliers.

6. REFERENCES

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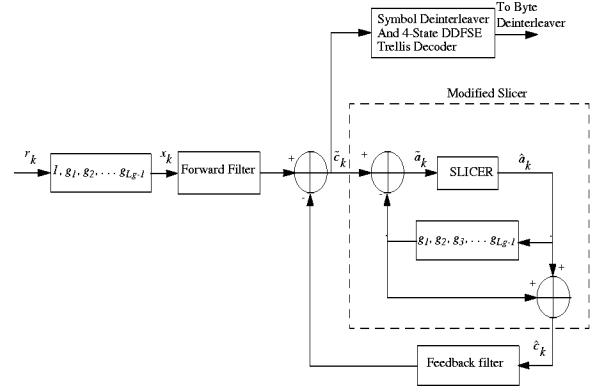


Figure 2: Rejection filter, DFE and trellis decoder configuration.

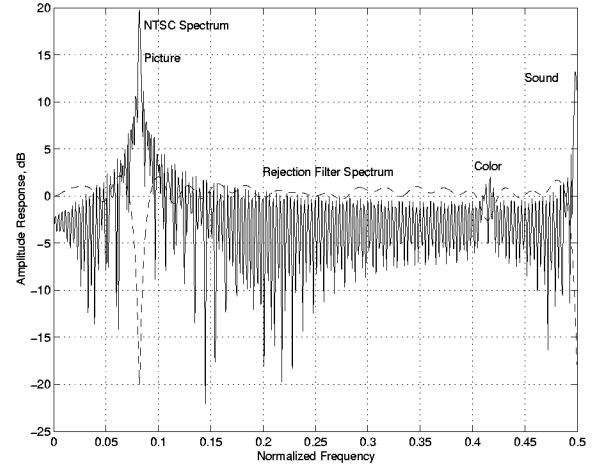


Figure 3: Spectra of co-channel NTSC and rejection filter

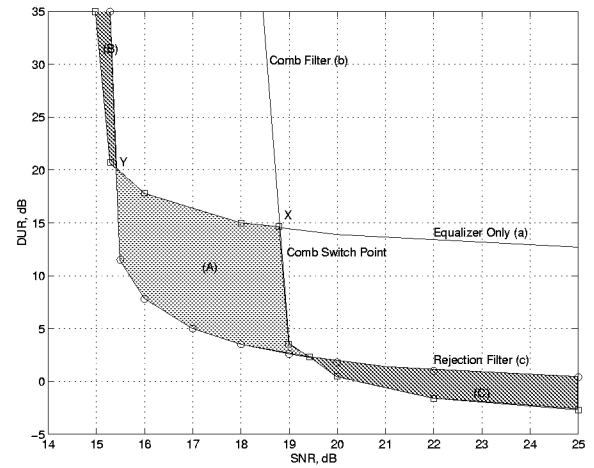


Figure 4: Performance of the comb filter and rejection filter systems.