Ghost Cancellation of Analog TV Signals: With Applications to IDTV, EDTV, and HDTV

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***Abstract*—Techniques that can cancel ghosts in received analog TV (for IDTV, EDTV, and HDTV) signals are presented. The fact that there are short periods of time without the analog signal (the horizontal flyback interval between the lines) is utilized to periodically cleanse a finite impulse response (FIR) or an infinite impulse reponse (IIR) equalizer. This line-by-line processing (cleansing) overcomes the limitations of standard equalizers to allow for 40–50 dB of suppression of ghosts, even with nulls in the spectrum, as long as the ghost delay is less than the period of time without the analog signal. Furthermore, by using time inversion in combination with line-by-line processing, the stability problem of the conventional IIR equalizer can be eliminated. It is shown that it may be possible to implement this IIR equalizer on a single digital integrated circuit. Alternatively, an FIR equalizer can be used which, although it requires multiple chips (i.e., more taps), can acquire and adapt to the ghosted channel more rapidly than an IIR equalizer. With line-by-line processing, FIR and IIR equalizers can eliminate any ghost with delays up to 11 μs in IDTV or EDTV, and any ghost with delays up to the length of the interval between each line without the analog signal in HDTV. For larger delays, we show how a standard IIR or FIR equalizer can be used as a preprocessor to eliminate small ghosts and an adaptive antenna can be used to eliminate large ghosts. Thus, using these techniques in combination with line-by-line processing, we can eliminate the ghosting problem in nearly all TV receivers.**

I. Introduction

MULTIPATH propagation is a significant source of picture quality degradation in television transmission. The ghosts in the received TV image can be a serious problem with NTSC signals and will be an even more important problem with improved definition TV (IDTV), extended definition TV (EDTV)[1], and high definition TV (HDTV) where these ghosts must be removed to realize the full quality improvement.

Numerous techniques [1]–[8] have been studied for the elimination of ghosts in TV images. For example, deghosting circuitry is commercially available (in Japan) for EDTV that can achieve from 20 to 30 dB of ghost suppression for ghosts up to 6 dB below the main signal level [2], [9]. However, all previously reported techniques are limited in the magnitude or type of ghosts that can be adequately suppressed with analog signals[2] (as in IDTV, EDTV, and most HDTV techniques). In particular, all these techniques fail when there is a null in the channel spectrum.[3] Specifically, with a null, a finite impulse response (FIR) equalizer cannot adequately suppress the ghosts while an infinite impulse repsonse (IIR) equalizer greatly enhances the noise in the picture. Such a null occurs with a single ghost (reflection) with the same magnitude as the desired signal (e.g., with a nonabsorbing reflector such as most buildings). A null can also occur with multiple ghosts even when they are weaker than the desired signal. Therefore, nulls in the channel spectrum (i.e., zeros in the z-domain that approach the unit circle) are a problem for many TV receivers [10]. Unfortunately, it is difficult for the viewer to determine the location of the zeros of the channel from the ghosted picture. Thus, not only do current techniques fail to adequately suppress ghosts in many cases (28% of the cases in one study [6]), the user may have to purchase[4] and test the deghosting electronics at home in order to determine it if will work properly.

In this paper, we present techniques that can eliminate ghosts in received analog TV signals in all of the cases described above. With our method, we utilize the fact that there are short periods of time without the analog signal (i.e., the horizontal flyback interval between the lines) to periodically cleanse (zero all the samples in the equalizer) an FIR or IIR equalizer. This periodic cleansing (line-by-line processing) eliminates the ghost suppression problem of a finite length FIR equalizer by shifting the ghost outside the line, and reduces the noise enhancement for the IIR equalizer. Thus line-by-line processing overcomes the limitations of conventional FIR and IIR equalizers to allow for 40–50 dB of suppression of ghosts, even with nulls in the spectrum—as long as the ghost delay is less than the period of time without the analog signal (e.g., the horizontal flyback interval for NTSC signals). We show that it may be possible to implement the IIR equalizer on a single digital integrated circuit (an enhanced version of a chip previously designed by S. Rao [11]). Alternatively, an FIR equalizer can be used which, although it may require multiple chips, can acquire and adapt to the ghosted channel more rapidly than an IIR equalizer. The technique can eliminate any ghost with delays up to 11.2 μs (the horizontal flyback interval in NTSC) for IDTV or EDTV (or the period without the analog signal in HDTV). For ghosts with larger delays, we show how a conventional IIR or FIR equalizer can be used as a preprocessor to eliminate the small ghosts and adaptive antennas can be used to eliminate the large ghosts. Thus, using these techniques in combination with line-by-line

[1] IDTV modifies the NTSC receiver for improved picture quality (e.g., by interpolation between the lines) using the present NTSC transmitted signal. EDTV also modifies the NTSC receiver for improved picture quality, but requires modification of the present NTSC transmission (e.g., by adding a training sequence for use in characterizing the multipath channel), although the signal is still compatible with NTSC receivers.

[2] Ghost elimination with digital signals (as in some proposed HDTV systems) generally requires a lower level of suppression and can be achieved using equalization techniques commonly used in the modem art, such as decision feedback equalization (DFE) [21].

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[3] That is, when there is a zero in the channel frequency transfer function, or equivalently a zero on the unit circle of the channel z-transform transfer function.

[4] The commercially available deghosting electronics retails for $750 as of fall 1989.

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processing, we can eliminate the ghosting problem in nearly all TV receivers.

Our techniques can be used with present NTSC signals to suppress all ghosts 20–30 dB below the main signal (IDTV). With better reference signals transmitted with NTSC (as in some EDTV proposals), suppression of all ghosts to 40–50 dB is possible. For HDTV, our techniques can be used to suppress the ghosts to 40–50 dB in the analog portion of the HDTV (i.e., hybrid analog/digital) signal. In these HDTV systems, the maximum delay for suppression of ghosts of any magnitude by line-by-line processing depends on the length of the interval without the analog signal between each line. This interval can be increased to eliminate the ghosting problem in the analog signal of HDTV systems.

In Section II we discuss the fundamental limits of ghost reduction techniques with continuous-time, analog signals. In Section III, we show these limits can be overcome in TV signals by line-by-line processing, utilizing the fact that there is a period without the analog signal between the lines. Experimental results and applications are presented in Section IV. Finally, a summary and conclusions are presented in Section V.

II. Fundamental Limits

A. Problem Definition

Here we consider the fundamental limits placed on a ghost canceler for the analog portion of the TV signal. For NTSC signals (i.e., for IDTV and EDTV), as well as for HDTV in the U.S. (where NTSC compatibility is required), the channel bandwidth is 6 MHz, although the analog video signal occupies only about 4.15 MHz.[5] The analog NTSC signal is a vestigial sideband (VSB) modulated signal. We consider two types of transmission with multipath propagation: cable and air. Multipath propagation in cable is caused by reflections from connectors, discontinuities, etc., in the cable.[6] With such reflections, the first signal received is the strongest, followed by much weaker signals (ghosts). These ghosts are referred to as postcursor ghosts and can usually be eliminated by simple, conventional ghost cancellation techniques, such as FIR equalization. In air transmission, ghosts are caused by reflections from buildings, trees, mountains, etc. If a line-of-sight exists between the transmit and receive antennas, then only postcursor ghosts will be present, although these ghosts may have signal levels approaching that of the main signal. If such a line-of-sight does not exist, then the strongest signal may arrive after the attenuated direct signal and these ghosts are called precursor ghosts; postcursor ghosts may also be present [6]. Such a situation is depicted in Fig. 1. In some cases, it may even be difficult to determine which is the strongest signal and this signal may itself be distorted since it can be a reflected signal. Also, the reflectors may be distributed, which results in a continuum of (or distorted) ghosts. Typically, however, the ghosts are few and widely-spaced [8] with delays ranging from -4 μs (precursor ghosts) to 37 μs [4], although the larger ghosts typically have smaller delays. The ghosts can change on the order of a second, with changing ghosts most commonly observed when indoor antennas are used [6], although flutter due to airplanes can also lead to ghosts with variations at this rate [4], [6]. In general, ghosts must be suppressed such that they are 40–50 dB below the main signal in order that they are not noticeable, while noise, because it is random, can be 10 dB higher than the ghosts before it is noticeable[7].

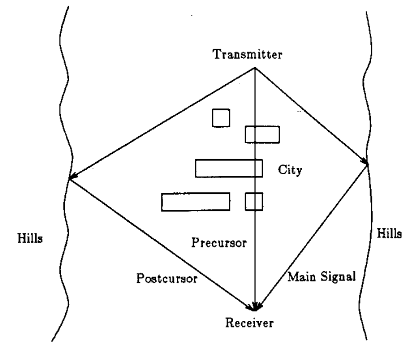

Fig. 1. Ghosting scenario where both precursor and postcursor ghosts are present.

The delay between the ghost and the main signal is given by the difference in path lengths, or

$$\tau = l/c, \tag{1}$$

where l is the difference in path lengths between the two signals and c is the speed of light (3×10^8 m/s). For example, an 11.2 μs delay corresponds to a 3.3 km path length difference, i.e., reflection from an object at least 1.6 km away. For the NTSC signal, the signal is transmitted in frames, one frame every 1/30 s, with 525 lines per frame. The line duration is 63.5 μs with 11.2 μs for the horizontal flyback interval. During this 11.2 μs interval, the analog signal is not transmitted—only completely predictable (deterministic) signals are transmitted. This will be used to our advantage in Section III. Also, during the vertical blanking interval of 1.27 ms, which occurs twice per frame (once per field), the analog signal is not transmitted.[8] With multipath propagation, the channel impulse in $h(t)$ is in general a continuous function of time. However, if we are only interested in reflectors that generate ghosts above some threshold, then we may be able to model the channel as a sum of I discrete reflectors

$$h_c(t) = \sum_{i=1}^{I} a_i \delta(t - \tau_i), \tag{2}$$

where a_i and τ_i are the amplitude and delay of the reflectors, and $\delta(t)$ is the Dirac delta function. The reflectors are usually distributed reflectors such as mountains and buildings, which limit the accuracy of this model and can result in distorted ghosts. However, since the signal is bandlimited, we can always (even with distributed reflectors, distorted ghosts, and ghosts with arbitrary delay [12], [13]) use a discrete channel model

$$h(t) = \sum_{i=1}^{J} b_i \delta(t - iT) \tag{3}$$

[5] Our results can easily be extended to analog TV signals with a wider bandwidth.

[6] Note that if the cable company receives its signal via an antenna, there may be a small amount of multipath propagation caused by this air transmission [2], [24].

[7] One novel method to reduce ghosting problems is to scramble the signal before transmission so that the ghosts appear as noise in the received picture [22]. However, with large ghosts, this noise is noticeable.

[8] During these intervals well-defined (deterministic) synchronization pulses are transmitted. Since these signals are known *a priori*, we can use these signals as a reference signal to determine the transfer function of the channel (see Section II-C-1). As long as the ghost duration is less than the horizontal flyback or vertical blanking interval, we can obtain a very accurate channel model.

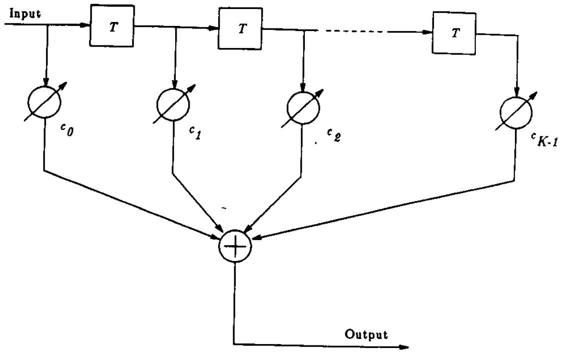

Fig. 2. Block diagram of tapped delay line equalizer (FIR filter) with K taps.

where the samples are taken at the Nyquist rate, $T = 1/2f_s$ and f_s is the signal bandwidth. Unfortunately, since we have analog signals, even with only one reflector, an accurate (accounting for ghosts up to 40-50 dB below the main signal) discrete channel model may require many samples J.

B. Conventional FIR and IIR Equalizers

Let us now consider the fundamental limits on ghost suppression of continuous analog signals. Ghost suppression at the receiver can be accomplished by passing the receiver signal through a filter with a transfer function that is the inverse of the channel.[9] Exact inversion of the channel of (3) can be obtained by an infinite impulse response (IIR) equalizer, while an approximate channel inverse can be achieved by a finite impulse response (FIR) equalizer. Below we determine the fundamental limits and advantages of these two equalizer architectures.

1. FIR: Fig. 2 shows a block diagram of a tapped delay line (FIR) equalizer with K taps with weights c_i, $i = 0, \cdots, K-1$. The delay between taps is $T = 1/2f_s$. Let us first consider the number of taps required to suppress a single ghost with a given input desired-to-undesired signal power ratio $(D/U)_{\text{in}}$, delay τ, and output desired-to-undesired signal ratio $(D/U)_{\text{out}}$. For the simple case where the ghost delay is a known integer multiple of the tap delay $\tau = kT$, the FIR equalizer that approximates the channel inverse is given by

$$h_{\text{FIR}}(t) = \delta(t) + \sum_{i=1}^{M-1} \left(-\sqrt{(D/U)_{\text{in}}}\right)^{-i} \delta(t - ikT), \quad (4)$$

i.e., the tap weights for the FIR equalizer are

$$c_0 = 1 \quad (5)$$

$$c_{ik} = \left(-\sqrt{(D/U)_{\text{in}}}\right)^{-i}, \qquad i = 1, \cdots, M-1, \quad (6)$$

$$c_j = 0, \qquad j \neq ik, \; i = 1, \cdots, M-1, \quad (7)$$

where the total number of taps in the equalizer is $K = k(M-1) + 1$. The output pulse (channel convolved with equalizer response) is

$$h(t)*h_{\text{FIR}}(t) = \delta(t) + \left(-\sqrt{(D/U)_{\text{in}}}\right)^{-M} \delta(t - MkT), \quad (8)$$

where * denotes convolution. Thus, after equalization the ghost is at $t = MkT$ with $(D/U)_{\text{out}} = (D/U)_{\text{in}}^M$, implying that $M = \log (D/U)_{\text{out}} / \log (D/U)_{\text{in}}$. In other words, $\log(D/U)_{\text{out}} / \log(D/U)_{\text{in}}$ nonzero taps are required, spaced at kT, to suppress a single ghost. For example, if $(D/U)_{\text{in}} = 5$ dB, then 10 taps are required for $(D/U)_{\text{out}} = 50$ dB. Thus, to suppress any single ghost with an unknown delay of any multiple of T up to kT, an equalizer with $K = k(M-1) + 1$ or $k(\log(D/U)_{\text{out}} / \log(D/U)_{\text{in}} - 1) + 1$ taps is required (although only M taps are active, i.e., have nonzero weights, with a single ghost). For example, to suppress any single ghost with $(D/U)_{\text{in}} < 5$ dB and $\tau < 11.2$ μs to $(D/U)_{\text{out}} > 50$ dB requires a 1212 tap equalizer (with $f_s = 6$ MHz, $T = 83.3$ ns).

However, the ghost may not always have delays that are integer multiples of T. As shown in Appendix I, for $\tau \neq kT$, up to N additional taps may be required around each of the taps required for a ghost at kT, where

$$N \leq \left((D/U)_{\text{out}}\right)^{1/6}. \quad (9)$$

Thus, for $(D/U)_{\text{out}} = 50$ dB, $N \leq 7$, and, to cancel a single ghost with $D/U_{\text{in}} = 5$ dB, up to $7 \cdot 10 = 70$ (active) taps may be required. Thus, all 1212 taps may have to be used (i.e., all 1212 tap weights could be nonzero) with only 18 ghosts.

The main point to note, however, is that for a fixed number of taps (as would be present with a given implementation), the ghost suppression decreases with ghost level (i.e., $(D/U)_{\text{out}} = (D/U)_{\text{in}}^M$). For example, with 1212 taps, $(D/U)_{\text{out}} = 30$ dB with $(D/U)_{\text{in}} = 3$ dB, and the level of suppression approaches zero as the ghost level approaches the desired signal level.

Finally, consider the multiple ghost case. Here, we simply note that the FIR equalizer requires an increasing number of taps as one or more of the zeros of $H(z) = Z\{h_k\}$, where Z denotes the z-transform and $h_k = h(kT)$, approach the unit circle. With multiple ghosts, the zeros of $H(z)$ can approach the unit circle if $\sum |a_i|$ approaches (or exceeds) the desired signal level [10], where the a_i's are the ghost amplitudes from (2). Thus, a fixed length FIR equalizer may not be able to adequately suppress ghosts even if the individual ghosts have powers several dB below the desired signal power.

2. IIR: Fig. 3 shows a feedback tapped delay line (IIR) equalizer with K taps with weights c_i, $i = 0, \cdots, K-1$. The delay between taps is $T = 1/2f_s$, as before. For a single ghost of any magnitude and with a delay of $\tau = kT$, channel inversion by the equalizer requires only one tap, independent of $(D/U)_{\text{in}}$ and $(D/U)_{\text{out}}$. Specifically, the tap weights for the IIR equalizer are

$$c_k = \left(-\sqrt{(D/U)_{\text{in}}}\right)^{-1}$$
$$c_j = 0 \qquad j \neq k. \quad (10)$$

Note that $(D/U)_{\text{out}} = \infty$, i.e., the ghost is completely canceled. With $\tau \neq kT$ (following the previous analysis), the number of taps required per ghost is given by (9), i.e., up to 7 taps may be required to suppress a single ghost to $(D/U)_{\text{out}} = 50$ dB.[10]

Note that the required number of taps for a given $(D/U)_{\text{out}}$ with the IIR equalizer does not increase with the ghost level, as with the FIR equalizer. Specifically, a K-tap IIR equalizer can

[9] Ghost suppression for air transmission can also be achieved by changing the received antenna pattern. This is considered further in Section III-C-2.

[10] Alternatively, we could oversample the ghosted signal to reduce the number of taps per ghost. In particular, for $(D/U)_{\text{out}} = 50$ dB, if we oversample 7 times (42×10^6 samples/s-corresponding to the 7 taps without oversampling), then only one tap per ghost is required, but this would require must faster (and costlier) A/D converters.

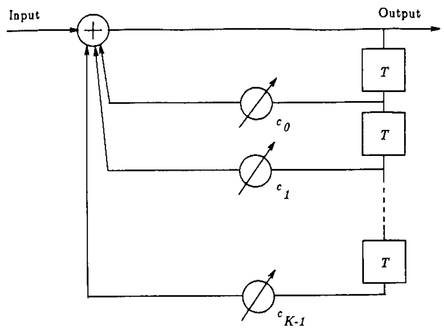

Fig. 3. Block diagram of feedback tapped delay line equalizer (IIR filter) with K taps.

always suppress, to a given $(D/U)_{\text{out}}$, any ghost (assuming $(D/U)_{\text{in}} \geq 1$ and $\tau > 0$, i.e., a weaker postcursor ghost) with $\tau < (K - (D/U)_{\text{out}}^{1/6})T$, where $(D/U)_{out}^{1/6}T$ is subtracted because up to $(D/U)_{\text{out}}^{1/6}$ taps are required when $\tau \neq kT$. For example, to suppress any ghosts within 11.2 μs, only 142 taps are required. However, although an IIR equalizer has the advantage over an FIR equalizer that it can always adequately suppress a ghost with a fixed number of taps, it has the disadvantages of greater noise enhancement and stability problems, as described below.

The noise enhancement of the equalizer is the ratio of the output noise power N_{out} to the input noise power N_{in}. For example, with a single ghost at $(D/U)_{\text{in}}$ and the (worst) case[11] of $\tau = T$,

$$\frac{N_{\text{out}}}{N_{\text{in}}} = \begin{cases} \dfrac{1 - (D/U)_{\text{in}}^{-M}}{1 - (D/U)_{\text{in}}^{-1}} & \text{FIR}, \\[2ex] \dfrac{1}{1 - (D/U)_{\text{in}}^{-1}} & \text{IIR}. \end{cases} \tag{11}$$

Thus, for the FIR with 1212 taps (and the IIR with 142 taps), the noise enhancement for $(D/U)_{\text{in}} = 6$, 3, and 1 dB is 3, 5, and 10 dB, respectively, for both the FIR and IIR. However, as $(D/U)_{\text{in}}$ approaches 0, the noise enhancement of the IIR approached infinity, while that of the FIR approaches 31 dB (with 1212 taps). Of course, with an FIR equalizer, the ghost suppression will not be adequate long before noise enhancement is a problem, and, thus, noise enhancement is not a concern. However, with an IIR equalizer, noise enhancement limits the maximum ghost level for which adequate picture quality can be obtained with equalization.[12]

Another problem is the stability of the IIR equalizer. With a single ghost, the IIR equalizer is stable only with postcursor ghosts—it is unstable with precursor ghosts, i.e., when the delayed signal is stronger than the first signal. Note that this is not a problem with an FIR equalizer, which, because it uses a feedforward architecture, is always stable. Specifically, the IIR equalizer is unstable when the zeros of the z-transform of the channel response are outside the unit circle. Note also that as the zeros approach the unit circle, noise enhancement degrades the picture quality.

With multiple ghosts, the performance of the IIR equalizer depends on the location of zeros (as with the single ghost case). Again, as the zeros approach the unit circle, noise enhancement degrades the picture quality, and the equalizer is unstable for zeros outside the unit circle, which, as discussed for the FIR equalizer, can occur even when the ghosts are several dB below the desired signal (without precursor ghosts).

C. Other Issues

1. Estimating the Channel: For the FIR equalizer, adaptation of the weights, to realize the channel inverse, can be accomplished using one of numerous techniques [14], with a suitable reference signal such as the horizontal or vertical sync. The vertical sync was used in early ghost cancellation schemes, but does not have sufficient bandwidth for accurate ghost suppression [4]. Thus, more complicated training signals are being considered that are placed in the vertical blanking interval (such as in [8] and GCR signal [2] used in Japan). Fig. 4 shows how adaptation of the weights can be achieved using the LMS algorithm [23], one of the simplest techniques and the one most widely used, particularly since it is easy to implement in VLSI. With the LMS technique, convergence to an accuracy of 40–50 dB generally requires the number of iterations on the order of 10 times the number of taps, or 10K. Thus, with one iteration per line (i.e., using the horizontal interval test signal), the convergence time is 635K μs, or 0.76 s with 1212 taps. This should be adequate for most TV receivers, where the ghosts may change within a second [4], [6]. However, with the LMS algorithm, the adaptation may become unstable if the ghosts change too rapidly. Indeed, rapidly varying ghosts are one of the main causes of poor ghost suppression in the commercially-available (in Japan) ghost suppression tuners [6]. Therefore, techniques to speed up the convergence, as, e.g., in [2] (also see [14]) may be required, particularly if a signal during the vertical blanking interval is used as the reference signal. However, even with the fastest algorithms, the adaptation speed will be limited by the signal-to-noise ratio and the frame rate (if the reference signal is only present during the vertical blanking interval).

For the IIR equalizer, adaptation of the weights can be carried out in a similar manner as that for the FIR equalizer, e.g., with the LMS algorithm as shown in Fig. 4, with the significant difference that the filter weights estimate the channel (not the channel inverse) and these weights are copied into the IIR structure shown in Fig. 4(c). Thus, for the IIR equalizer, the reference signal and received signals are reversed so that the tapped delay line of Fig. 4 is the channel model $H(z)$, trying to approximate the channel $H_c(z)$ (rather than $H_{FIR}(z)$, trying to approximate the inverse model $1/H(z)$ for the FIR equalizer). For the IIR equalizer, the channel is inverted by the feedback transfer function $1 - H(z)$. Thus, the weights in the IIR equalizer (c_i, $i = 1, \cdots, K-1$) are just the negative of the coefficients b_i in the channel model (3).

If the maximum ghost delay is greater than 11.2 μs, then the analog signal will interfere with the horizontal interval test signal, and an accurate channel model may be difficult to determine. In this case, the vertical interval test signal or other

[11] Note that we can decrease the noise enhancement by using weights that minimize the mean-squared-error due to both the noise and the ghosts, e.g., $N_{\text{out}} + U$. However, since the ghosts are more noticeable than the noise, a better criterion may be to minimize the mean-squared-error of $\alpha N_{\text{out}} + U$, $\alpha \ll 1$ (determined subjectively), which would result in about the same noise enhancement as (11).

[12] As discussed before, we can decrease the noise enhancement and thereby improve on this limit by using weights that minimize the ghosts plus noise $D/(U + N_{\text{out}})$, in the picture, rather than just the ghosts. Even further improvement could be obtained by using weights that minimize the sum of the ghosts and a fraction of the noise $D/(U + \alpha N_{\text{out}})$, since noise is less noticeable than ghosts. However, the improvement in either case may be marginal.

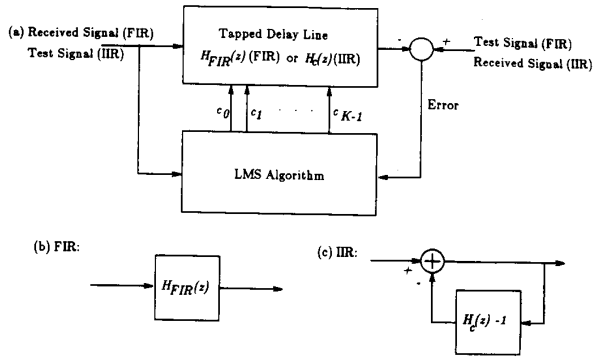

Fig. 4. Block diagram of weight adaptation for FIR and IIR equalizers.

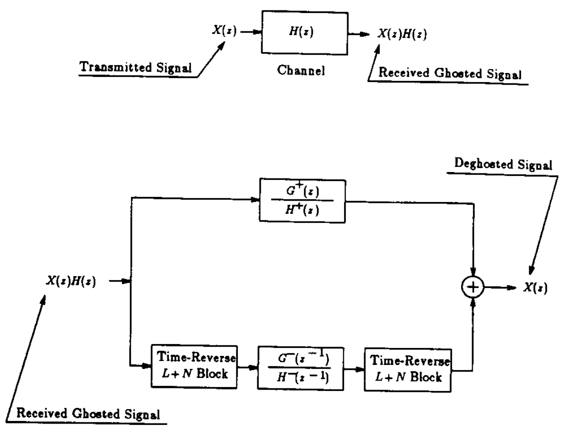

Fig. 5. Block diagram of realization of IIR filter with unstable modes using time-reversal.

training signal can be used to generate the reference signal. However, as discussed above, the adaptation time is increased by a factor of the number of lines (i.e., 525, or 262.3, since there is a vertical blanking interval in each field), which may be unacceptable, unless faster convergence algorithms are used [14].

2. Accuracy Of Weight Values: Next, consider the required accuracy of the equalizer to keep the quantization noise below a given level. Since the quantization noise appears random rather than coherent as ghosts, the TV viewer can tolerate a higher level of quantization noise—typically a noise level of 40 dB is not noticeable. For a K tap equalizer with n bits of accuracy in the tap weights, the signal-to-quantization-noise ratio S/N is

$$S/N = \frac{2^{-2n}}{12}K. \tag{12}$$

Thus, for a 40–50 dB S/N and 1212 taps, 10–12 bits of accuracy are required, while for 142 taps, 9–11 bits of accuracy are required.[13] Since most analog devices have only up to 8 bits of accuracy, digital representation and signal processing (i.e., digital chips) are required to obtain 40–50 dB signal-to-quantization noise ratio.

3. Implementation: Implementation of the algorithm requires a digital signal processor with 12 bits of accuracy of $2f_sK$ complex multiplications and additions per second. With $f_s = 6$ MHz and $K = 142$, 1.7×10^9 complex multiplications and additions per second are required. To determine the required number of chips for implementation, consider the tapped delay line chip by S. Rao [11], which, in its current version, is capable of 360×10^6 complex multiplications and additions per second on 22 mm^2 of active area. Thus, a chip is required for every 30 taps, or 5 chips for 142 taps. However, we can increase the capacity of the chip by increasing the active area to 105 mm^2. This would result in a chip with about 5 times more capacity, i.e., one chip for every 143 taps (which span 11.9 μs). Thus, the entire IIR equalizer could be implemented on a single digital chip. Alternatively, the FIR equalizer with 1212 taps would require 9 chips. Of course, if we only want to suppress fewer than 18 ghosts (see Section II-B-1), then fewer chips would be required.

4. Summary: In summary, the IIR equalizer has the advantage over the FIR equalizer that it requires taps only over the interval of ghost delays to suppress from one up to an arbitrary number of ghosts, while the FIR equalizer requires an increasing number of taps as $(D/U)_{\text{in}}$ approaches 0 dB (i.e., as the zeros of the channel response approach the unit circle). Thus, the IIR equalizer can be implemented with fewer chips (possibly only one). However, as $(D/U)_{\text{in}}$ approaches 0-dB, noise enhancement with the IIR equalizer degrades picture quality. The FIR equalizer, on the other hand, has the advantage that it can operate even when the zeros are outside the unit circle (e.g., with precursor ghosts), while the conventional IIR equalizer is unstable in this case.

Thus, both the FIR and IIR equalizers have unsatisfactory performance (ghost suppression or noise enhancement) as the zeros of the channel transfer function approach the unit circle. As discussed earlier, this is undesirable since many TV receivers may have ghosts of this type that, therefore, cannot be canceled by the above techniques, yet the user could not determine whether the equalizer would work without testing it at home. For satisfactory performance, the ghost suppression of the FIR equalizer must be improved, the noise enhancement of the IIR equalizer must be reduced, and the IIR equalizer must be made stable, under all ghosting scenarios.

III. Line-By-Line Processing

In Section II we assumed that the TV signal was a continuous analog signal; however, the NTSC TV signal has a dead time during the horizontal flyback interval. If the ghost duration is less than the horizontal flyback interval (11.2 μs or a 3.3-km path length difference)[14], then there is no interference between lines of the analog signal (deterministic signals during the horizontal flyback interval cause known interference which can be subtracted out, as discussed below) and line-by-line processing can be done. This eliminates the ghost suppression problem of the FIR equalizer (since the residual ghost appears outside the analog signal portion of the line), reduces noise enhancement in the IIR equalizer, and can be used to guarantee stability of the IIR equalizer, under all ghosting scenarios, as shown below.

A. FIR Equalizer

Here, we restrict our attention to the case where the total ghost delay variation is less than the horizontal flyback interval.

[13] While this accuracy is needed for the weights used in the FIR or IIR equalizer, additional accuracy is required in calculating these weight values for the channel model.

[14] In formats other than NTSC (such as some proposed HDTV formats), the period without the analog signal may be different than 11.2 μs. In this case, the maximum ghost delay for line-by-line processing will change accordingly. Here, we restrict our discussion to NTSC signals.

That is, the maximum precursor delay τ_{pre} plus the maximum postcursor delay τ_{post} is less than the horizontal flyback interval. In this case, all the information needed to determine a line is contained in the samples from τ_{pre} before to τ_{post} after the line, and no other nondeterministic signals are in these samples. Of course, any deterministic signals in the horizontal flyback interval are in these samples, but since these signals are known *a priori* (or their exact received level can easily be determined), they can be subtracted out. This is discussed in detail in Section III-C-1.

Thus, to deghost one line, we only need to use samples over that line (63.5 μs or 763 samples for a 12 MHz sampling rate). Therefore, the FIR equalizer needs to have a maximum length of $2 \times 52.3 + 11.2 = 115.8$ μs (since, to generate an output sample at a given time, the equalizer combines samples from up to 52.3 μs before to 52.3 μs after that time, plus the 11.2 μs delay variation) or 1398 taps to deghost the line[15]. In practice, we would simply zero the samples in the FIR equalizer (i.e., cleanse the equalizer) at the beginning of each line. Although the equalizer has 1398 taps (i.e., stored tap weights), since there are only 763 samples per NTSC line, only a maximum of 763 taps are active at any one time. Also, the FIR equalizer needs to generate an output only for each sample in the output picture, i.e., 52.3/63.5 of each line. Thus, the FIR has an average processing rate of a $763 \times 52.3/63.54 = 628$ tap equalizer, which requires only five of the previously mentioned chips. Note also that with line-by-line processing, as long as the total ghost delay spread is less than 11.2 μs, the ghosts are completely eliminated, for any ghost magnitude or location of zeros.

However, with the ghosts eliminated, we must now be concerned with the noise enhancement of the FIR equalizer. With 763 samples, the noise enhancement of the FIR equalizer will at most be 29 dB. Of course, this only occurs when all the taps are of equal magnitude (one ghost with $(D/U)_{\text{in}} = 0$ dB and $\tau = T$), which is unlikely. In all other cases, the noise enhancement will be less. Thus, with line-by-line processing, an FIR equalizer provides a good solution to the ghost cancellation problem, even though it requires more taps than an IIR equalizer.

B. IIR Equalizer

1. Zeros Inside the Unit Circle: For the IIR equalizer, line-by-line processing has the advantages that it reduces the noise enhancement and can be used to ensure stability. From the above discussion for the FIR equalizer, the reason for the reduction in noise enhancement should be clear. Since, with line-by-line processing, the IIR equalizer is cleansed at the end of each line, the noise enhancement is limited (for zeros inside the unit circle, i.e., a stable conventional IIR equalizer) to that of the FIR equalizer. Specifically, in the worst case, the noise enhancement is 29 dB rather than infinity.

2. Arbitrary Zeros: With line-by-line processing, the IIR equalizer is, of course, always stable. However, with zeros outside the unit circle, the noise enhancement can easily be very large (and exceed 29 dB), and the signal levels in the equalizer can saturate the devices in any given implementation. To avoid these problems, we propose the use of spectral factorization, along with time inversion. Specifically, we first factor the channel response $H(z)$ into the product of two polynomials—one with zeros inside ($H^+(z)$) and one with zeros outside ($H^-(z)$) the unit circle. Then, if we pass the received signal through $G^+(z)/H^+(z)$ and pass the time-reversed received signal through $G^-(z)/H^-(z)$, such that $G^+(z)/H^+(z) + G^-(z)/H^-(z) = 1/H(z)$, we can equalize the channel with minimal noise enhancement (29-dB noise enhancement in the worst case). Appendix II provides the details of this scheme (see Fig. 5). A similar technique was proposed in [10] (also see [12]).

Thus, with line-by-line processing, the IIR equalizer has the same performance (ghost suppression and noise enhancement) as the FIR equalizer, but requires fewer taps (142 versus 628 active taps).

The disadvantage of this technique, however, is that the factorization of the channel inverse into two separate polynomials can be complicated and time consuming. For example, consider the simple channel filter $\alpha_1 + z^{-1} + \alpha_2 z^{-2}$, which corresponds to one precursor and one postcursor ghost. Depending on the values of α_1 and α_2, we have the following possible situations for the locations of the zeros of this polynomial

α_1	α_2	location of zeros
0.7	0.5	both inside unit circle
0.5	0.7	both outside unit circle
0.5	0.5	both on unit circle
0.4	0.4	one inside, one outside unit circle.

As can be seen from this example, even with just two ghosts, the location of the zeros relative to the unit circle is not obvious. Thus, to factor the channel response into $H^+(z)$ and $H^-(z)$, we must find all the zeros of $H(z)$. Unfortunately, the factorization of a 142 degree polynomial is complicated and time consuming. For example, this factorization takes 1.5 min using the algorithm of [15] on a SUN 3/50 with the 68881 floating point coprocessor. Therefore, the adaptation time of the IIR equalizer will be longer than that of the FIR equalizer, although adaptation is still feasible (but may result in inadequate ghost suppression with rapidly changing ghosts).

C. Other Issues

1. Deterministic Signals in the Horizontal Flyback Interval: As noted in Section III-A although each line contains only samples from the ghosted analog signal of that line and not from the analog signal of other lines (as long as the total ghost delay variation is less than the horizontal flyback interval), signals in the horizontal flyback interval of the transmitted signal (such as the horizontal sync pulse) may be ghosted such that they have samples in two adjacent lines. Thus, line-by-line processing may cause these signals to interfere with the deghosted analog signal. However, as long as signals during the horizontal flyback interval are deterministic signals, e.g., digital signals or known waveforms, their shape and level without ghosting is known *a priori*. Since the channel response is also known, we can determine the sampled values for these signals and subtract them from the analog signal before equalization (this was also discussed in [4]). Thus, for line-by-line processing to adequately equalize the received analog signal, the only requirement is a period without any nondeterministic signals between the lines, with the ghost duration less than this period.

2. Ghosts with Larger Delays: As noted above, for line-by-line processing to work, the ghost delay must be less than 11.2 μs (for NTSC). Unfortunately, ghost delays as large as 24 μs (corresponding to a path length difference of 7.2 km) [16] are common and ghost delays up to 37 μs can occur [4]. However,

[15] If only postcursor ghosts are present (as with cable transmission, see Section II-A), then the FIR equalizer only needs to have a length of 52.3 μs or 628 taps.

line-by-line processing is only needed with large ghosts (or zeros near the unit circle), otherwise, conventional equalization techniques are adequate. Fortunately, since ghosts with large delay tend to be weaker (since they are reflected from more distant objects), conventional equalization techniques should be adequate for large delay ghosts. Specifically, two equalizers would be used, one for nearby and one for more distant ghosts, which is similar to the technique discussed in [7], where an FIR and an IIR filter were used. If any significant ghosts (i.e., reference signal energy) outside of 11.2 μs were detected, then a channel model outside of 11.2 μs could be determined, e.g., by calculating the correlation of the reference signal with the received signal (standard techniques, e.g., [2], can be used). As in [7], a standard IIR filter with a few small taps (assuming a few, weak ghosts with large delay) can be used to suppress these ghosts. The resulting signal after this preprocessor may only have significant ghosts within 11.2 μs, which can then be suppressed using line-by-line processing. However, if there are large ghosts within 11.2 μs, the IIR filter that uses the channel model outside of 11.2 μs may not adequately suppress all ghosts outside of 11.2 μs. In this case, the channel model (polynomial) over the entire delay range must be factored into two polynomials, one with zeros near the unit circle $H_l(z)$ (associated with large ghosts) and one with zeros near the origin $H_s(z)$. As before, a standard IIR filter can be used to compensate for $H_s(z)$ (i.e., the IIR filter has the transfer function $H_s^{-1}(z)$), followed by line-by-line processing to compensate for $H_l(z)$. Thus, in either case, by preprocessing the received signal, the long delay ghosts can be suppressed and line-by-line processing will work satisfactorily.

For large ghosts with large delays, we consider the following solution. Since such ghost are generated by large, distant objects, these reflectors will occupy only a small spatial angle from the point of view of the receiving antenna. In general, such ghosts can be reduced in magnitude by adjusting the receive antenna so that its pattern is very weak in the direction of the main reflections. Since the location of these large objects (such as buildings or mountains) usually would be fixed, only a one time adjustment would be required. Simple rabbit ears can be adjusted to suppress strong ghosts. Indeed, this is the earliest method of ghost suppression [13], and is the method recommended for suppression of large ghosts so that standard FIR or IIR filters give adequate ghost suppression [4]. In severe cases, adaptive antennas can be used to suppress the large ghosts [17], [18]. With adaptive antennas, the signals from two or more antennas are weighted and combined to maximize the signal-to-noise plus interference (ghosts) power ratio. A commonly used technique for weight adaptation is the LMS algorithm (used in a similar manner to Fig. 4), whereby the weights are adjusted to minimize the mean-squared-error of the difference between the reference signal and the received reference signal. Adaptive antennas can be used in combination with line-by-line processing, and even preprocessors using conventional FIR or IIR equalizers, to eliminate all types of ghosts. However, the main remaining concern is that the adaptation speed must be faster than the rate of change of the ghosts - an adaptation speed under 1 s is generally considered adequate to track nearly all ghosts [4], [6].

IV. Examples and Applications

A. Examples

We now consider three examples that illustrate the three advantages of line-by-line processing: 1) elimination of the ghost suppression limitation of a conventional FIR equalizer, 2) reduction of the noise enhancement limitation of a conventional IIR equalizer, 3) usage of an IIR with zeros inside and outside the unit circle. These advantages can be demonstrated on a large TV (e.g., 25 inch) with moving images, where the picture degradation with $S/N < 40$ dB and $(D/U)_{out} < 40–50$ dB is noticeable. However, with still photos of paper quality, the picture degradation isn't noticeable unless the S/N and $(D/U)_{out}$ are much lower (e.g., $S/N < 20$ dB, $(D/U)_{out} < 10$ dB). Thus, in the following examples, the noise and ghost levels are much higher than would typically be present in TV receivers, although these advantages also hold at lower noise and ghost levels.

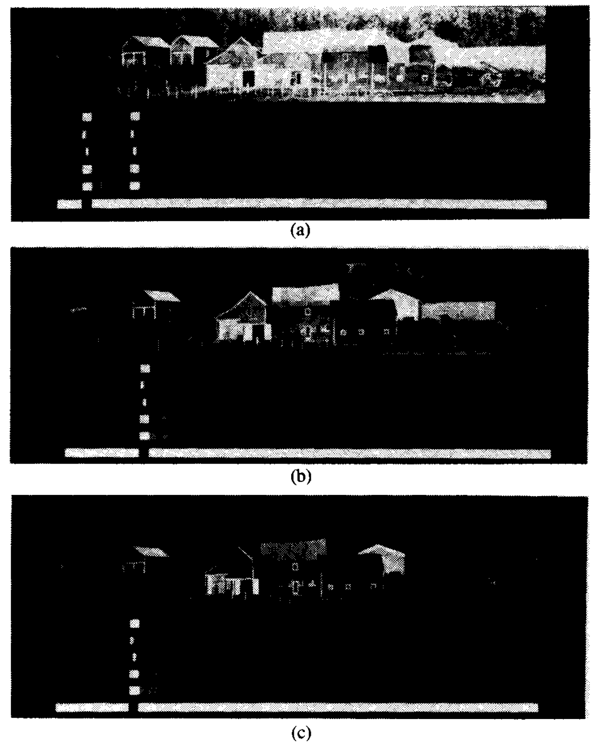

Fig. 6. Suppression of single ghost with $(D/U)_{in} = 1$ dB and $\tau = 5$ μs by an FIR equalizer. (a) Original image. (b) Image deghosted by conventional FIR equalizer. (c) Image deghosted by FIR equalizer using line-by-line processing. (For color supplement see page 163.)

First, consider a single ghost with $(D/U)_{in} = 1$ dB and $\tau = 5$ μs, as illustrated in Fig. 6(a). Thus,

$$h(t) = \delta(t) + 0.8\delta(t - 5\ \mu\text{s}). \quad (13)$$

For a conventional FIR equalizer with 628 taps, from (4),

$$h_{\text{FIR}}(t) = \delta(t) + \sum_{i=1}^{10} (-0.8)^{-i}\delta(t - 5i\ \mu\text{s}), \quad (14)$$

i.e., there are 11 active taps. The channel plus equalizer response is then from (8):

$$\begin{aligned} h(t)*h_{\text{FIR}}(t) &= \delta(t) + (-0.8)^{11}\delta(t - 55\ \mu\text{s}), \\ &= \delta(t) + (-0.09)\delta(t - 55\ \mu\text{s}). \quad (15) \end{aligned}$$

Therefore, after equalization, the ghost is at $\tau = 55$ μs with $(D/U)_{out} = 11$ dB. Since the line length is 63.5 μs, the ghost actually appears as a precursor ghost with $\tau = -8.5$ μs, as

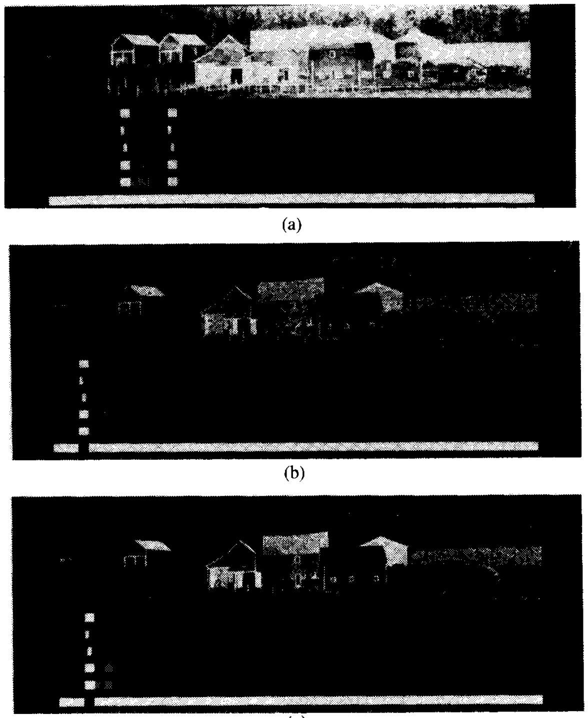

Fig. 7. Suppression of single ghost with $(D/U)_{in} = 0.18$ dB and $\tau = 5$ μs by IIR equalizer. (a) Original image. (b) Image deghosted by conventional IIR equalizer. (c) Image deghosted by IIR equalizer using line-by-line processing. (For color supplement see page 163.)

shown in Fig. 6(b). With line-by-line processing, however, the ghost is completely suppressed, as shown in Fig. 6(c).

Next, consider a single ghost with $(D/U)_{in} = 0.18$ dB and $\tau = 5$ μs, and a received maximum $S/N = 20$ dB (i.e., the signal power at saturation to noise power ratio) as shown in Fig. 7(a). With a conventional IIR equalizer with 142 taps, the noise enhancement is, from (11),

$$\frac{N_{out}}{N_{in}} = \frac{1}{1 - 0.96} = 25 \tag{16}$$

or the S/N in the equalizer output is 6 dB, as shown in Fig. 7(b). With line-by-line processing, the noise enhancement is, from (11), that of an FIR equalizer, or,

$$\frac{N_{out}}{N_{in}} = \frac{1 - (0.96)^{10}}{1 - 0.96} = 8.4 \tag{17}$$

or the worst S/N is the equalizer output is 11 dB, as shown in Fig. 7(c). Note that the S/N decreases (the noise increases) across the line from 20 dB to 11 dB.

Finally, consider two ghosts with $(D/U)_{in} = 8$ dB and $\tau = \pm 1.5$ μs, as shown in Fig. 8(a). Thus,

$$h(t) = \delta(t) + 0.4\delta(t - 1.5\ \mu s) + 0.4\delta(t + 1.5\ \mu s), \tag{18}$$

and

$$H(z) = 0.4z^{30} + 1 + 0.4z^{-30} = 0.2(2 + z^{-30})(2 + z^{30}), \tag{19}$$

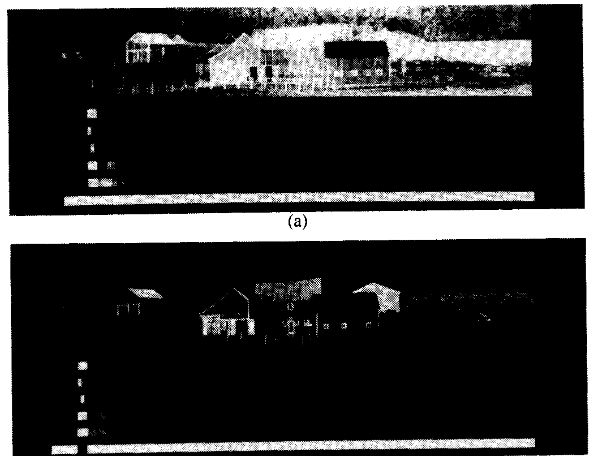

Fig. 8. Suppression of two ghosts with $(D/U)_{in} = 8$ dB and $\tau = \pm 1.5$ μs. (a) Original image. (b) Image deghosted by IIR equalizer using line-by-line processing with time inversion. (For color supplement see page 163.)

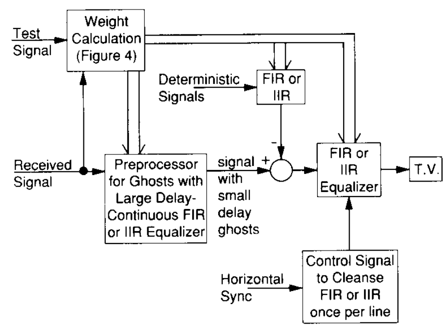

Fig. 9. Block diagram of ghost canceler.

and, therefore, half the poles are inside and half the poles are outside the unit circle. Fig. 8(b) shows the deghosted image when line-by-line processing with time inversion is used to equalize the channel.

B. Applications

Fig. 9 shows a block diagram of the ghost cancellation technique incorporating the major features discussed previously. Below, we consider applications of this technique for NTSC and HDTV receivers.

For conventional NTSC receivers, our techniques can be implemented as a box, connected between the antenna or cable input to the TV (set-top ghost cancellation tuners are commercially available in Japan [2], [6]). The box consists of an A/D, one or more digital chips, and a D/A (plus required frequency translation circuitry). For IDTV or EDTV, since digital signals are used within the TV, our techniques can be implemented by simply incorporating the digital chip(s) in the TV.

For HDTV, the application and need for our techniques depends on the HDTV format. If the HDTV format consists only of digital signals, then our techniques may not be needed, since conventional equalization techniques such as DFE can be used. However, since most HDTV proposals use a mixture of digital and analog signals, our techniques would be advantageous of deghosting the analog portion of the signal in all ghosting scenarios. In particular, most existing HDTV systems, such as the MUSE system [19] in Japan and MAC system in Europe, use analog video encoding with sub-Nyquist rate sampling [8]. Although these systems use digital processing which eliminates the need for horizontal flyback intervals, clock synchronization is critical, and therefore synchronization pulses are periodically (once per line in MUSE) inserted into the signal. Since these are known signals transmitted without the analog signal, line-by-line processing can be used, with the allowable ghost delay depending on the duration of the synchronization pulses.

For the spectrum-compatible HDTV systems under study in the U.S., the NTSC signal can be deghosted by line-by-line processing as before, while the augmented signal, which typically contains analog samples and synchronization pulses (as in MUSE) can be deghosted by line-by-line processing, again with the allowable ghost delay depending on the duration of the synchronization pulses. For example, in the Zenith HDTV proposal [20], synchronization pulses are sent once per line with a duration of about 2 μs under consideration. Finally, we note that simple changes in the HDTV signals may be possible that could allow for longer periods without the analog signal and thereby allow our techniques to eliminate large ghosts with even longer delays in these systems.

V. Conclusion

In this paper, we discussed the ghost problem for analog TV, and showed the limitations of conventional FIR and IIR equalizers. Specifically, as the zeros of the channel response approach the unit circle, a fixed length FIR equalizer cannot adequately suppress the ghosts, while an IIR equalizer will have too much noise enhancement and become unstable. We then showed how by using line-by-line processing with time inversion these limits can be avoided, using an FIR or IIR equalizer to adequately suppress ghosts without significant noise enhancement in all cases, as long as the ghost duration was less than the period without the analog signal—11.2 μs in NTSC. The IIR equalizer may be implemented on only one digital chip, while the FIR equalizer requires four times the processing capability. However, adaptation of the IIR equalizer is more complicated and time consuming because the channel polynomial must be factored. For ghosts with larger delays, we showed how a conventional IIR or FIR equalizer can be used as a preprocessor to eliminate the small ghosts and adaptive antennas can be used to eliminate the large ghosts. Thus, with these techniques, the ghosting problem can be eliminated in nearly all TV receivers.

Appendix I
Number of Taps for Ghosts with Arbitrary Delay

Let the impulse response of the transmitting filter be $g(t)$ and be bandlimited to f_s. With the channel as given by (3), the impulse response of the transmitter and channel is

$$h(t) = g(t) + \frac{1}{\sqrt{(D/U)_{\text{in}}}} g(t-\tau). \tag{20}$$

For illustrative purposes, we will assume $|\tau| < 1/2f_s = T$. To cancel the ghost, the equalizer must subtract $1/\sqrt{(D/U)}_{\text{in}} g(t-\tau)$ from the received signal, but since the tap delays are spaced at $T\ (= 1/2f_s)$, it is required that

$$g(t-\tau) = \sum_{l=-\infty}^{\infty} g(t_1 - \tau)\, \text{sinc}\,(2f_s(t-t_l)). \tag{21}$$

If the summation is truncated at L samples (i.e., $N = 2L+1$), the error at $t = \tau$ will be approximately

$$\epsilon = 2 \sum_{l=L+1}^{\infty} g(t_l - \tau)\, \text{sinc}\,(2f_s(\tau - t_l)). \tag{22}$$

For the worst case, let $g(t)$ be a sinc function (i.e., the transmitter filter is a brick-wall filter). Then,

$$\epsilon = 2 \sum_{l=L+1}^{\infty} \text{sinc}^2\,(2f_s(\tau - t_l)) \tag{23}$$

$$= 2\,\text{sinc}^2\,(2f_s\tau) \sum_{l=L+1}^{\infty} \left(1 - \frac{l}{2f_s\tau}\right)^{-2}, \tag{24}$$

which, satisfactorily, vanishes for integer $2f_s\tau$. Keeping $|2f_s\tau| < 1$, (24) becomes

$$\epsilon = \frac{2\sin^2(2\pi f_s\tau)}{\pi^2} \sum_{i=L+1}^{\infty} (l - 2f_s\tau)^{-2}, \tag{25}$$

For $l \gg 1$, the sum is excellently approximated by

$$\sum_{i=L+1}^{\infty} l^{-2} \approx \frac{1}{L}. \tag{26}$$

Thus,

$$\epsilon \approx \frac{2}{\pi^2 L} \tag{27}$$

or for $N = 2L+1$, since

$$\epsilon = \left(\frac{U_{\text{out}}}{U_{\text{in}}}\right)^{1/2} = \left(\frac{(D/U)_{\text{in}}}{(D/U)_{\text{out}}}\right)^{1/2},$$

$$N = \frac{4}{\pi^2}\left(\frac{(D/U)_{\text{out}}}{(D/U)_{\text{in}}}\right)^{1/2} + 1. \tag{28}$$

If we assume that the equalizer uses the strongest signal as the desired signal, then $(D/U)_{\text{in}} \geq 1$, and, from (28),

$$N \leq \frac{4}{\pi^2}((D/U)_{\text{out}})^{1/2} + 1. \tag{29}$$

Thus, for $(D/U)_{\text{out}} = 50$ dB $N \leq 130$. In practice the transmitting filter will not cutoff sharply at f_s. That means that for large l, $g(t_l - \tau)$ will fall off with l more rapidly than l^{-1} (as with the sinc function). For TV signals (although filters will vary), it is reasonable to assume a l^{-3} type behavior. Then we have, instead of $\sum l^{-2}$, $\sum l^{-4} \approx L^{-3}$, and, from (29),

$$N \leq ((D/U)_{\text{out}})^{1/6}. \tag{30}$$

Thus, for $(D/U)_{\text{out}} = 50$ dB, $N \leq 7$.

Appendix II
IIR For Arbitrary Pole Location

Consider the simple case $H(z) = \alpha_1 + z^{-1}$, $0 < |\alpha_1| < 1$. Let $X(z)$ be the z-transform of one transmitted block $\{x_0, x_1, \cdots, x_{L-1}\}$. We have the corrupted received signal with z-transform:

$$X(z)H(z) = (x_0 + x_1 z^{-1} + \cdots + x_{L-1} z^{-L+1}) \cdot (\alpha_1 + z^{-1}) \tag{31}$$

$$= x_0\alpha_1 + (x_0 + x_1\alpha_1)z^{-1} + \cdots + x_{L-1}z^{-L} \tag{32}$$

from which we would like to extract $X(z)$. Processing $X(z)H(z)$ by $1/H(z)$ would have the desired effect, but $1/H(z)$ is an unstable filter. Although due to the block-by-block processing, stability is not a problem for the actual signal (we only look at the output for one block), any noise in the received signal will be amplified and may corrupt the output. We can, however, observe that in convolving two time sequences the two sequences are moved in opposite directions to calculate the consecutive output values, from beginning to end. One could also start from the end of the output time sequence and move towards its beginning. To accomplish this, both the input and the filter sequences need to be time-reversed, convolved, and the output time-reversed again so that the beginning becomes the end and vice versa. For the example above, when we reverse the received signal in time, we obtain

$$x_{L-1} + (x_{L-2} + x_{L-1}\alpha_1)z^{-1} + \cdots + x_0\alpha_1 z^{-L} \quad (33)$$

and process it by the time-reversed (and therefore stable) IIR filter

$$\frac{1}{1 + \alpha_1 z^{-1}} \quad (34)$$

It can easily be verified by long division that this process results in

$$x_{L-1} + x_{L-2}z^{-1} + \cdots + x_0 z^{-L+1}, \quad (35)$$

which, when time-reversed, yields

$$X(z) = x_0 + x_1 z^{-1} + \cdots + x_L z^{-L}. \quad (36)$$

In general, we have a channel filter $H(z)$ with zeros both inside and outside the unit circle. Let the total number of zeros $H(z)$ be N. $H(z)$ can be factorized as

$$H(z) = H^+(z)H^-(z) \quad (37)$$

such that $H^+(z)$ has all of its zeros inside the unit circle, the $H^-(z)$ has all of its zeros outside the unit circle. Then, $1/H(z)$ can be expanded as

$$\frac{1}{H(z)} = \frac{G^+(z)}{H^+(z)} + \frac{G^-(z)}{H^-(z)}. \quad (38)$$

Observe from (34) and (35) that time reversing a sequence of length L has the following effect on the z-transform of the time-reversed sequence:

$$X_R(z) = z^{-L+1}X(z^{-1}) \quad (39)$$

where $X(z)$ is the z-transform of the original sequence and $X_R(z)$ is the z-transform of the time reversed sequence. Observe, also, that $H^-(z^{-1})$ has all of its zeros inside the unit circle. In other words, $G^-(z^{-1})/H^-(z^{-1})$ is a stable filter. If we time-reverse the received signal block (of length $L + N$) which is corrupted due to $H(z)$, and process it with the stable filter $G^-(z^{-1})/H^-(z^{-1})$, we have the output

$$z^{-L-N+1}X(z^{-1})H(z^{-1})\frac{G^-(z^{-1})}{H^-(z^{-1})}, \quad (40)$$

which, when time reversed as a block of $L + N$ samples, yields

$$z^{-L-N+1}\left[z^{-L-N+1}X(z^{-1})H(z^{-1})\frac{G^-(z^{-1})}{H^-(z^{-1})}\right]_{z\leftarrow z^{-1}} = X(z)H(z)\frac{G^-(z)}{H^-(z)}. \quad (41)$$

By processing the received signal with the stable filter $G^+(z)/H^+(z)$, and adding to the right hand side of (41), we obtain the desired signal:

$$X(z)H(z)\left[\frac{G^+(z)}{H^+(z)} + \frac{G^-(z)}{H^-(z)}\right] = X(z). \quad (42)$$

The block diagram of the system is shown in Fig. 5.

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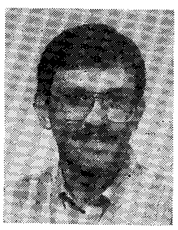

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