TRAINING SIGNAL AND RECEIVER DESIGN FOR MULTI-PATH CHANNEL CHARACTERIZATION FOR TV BROADCASTING

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ABSTRACT

The problem of multi-path propagation in television terrestrial broadcasting can be solved by channel equalization. The use of such equalization is needed for an enhanced NTSC system and is particularly critical in most HDTV proposals. We propose techniques for precisely characterizing a multi-path channel. These characterizations can be used to reduce hardware complexity and to speed up equalizer convergence.

I. INTRODUCTION

In an NTSC broadcasting environment, ghosted pictures are the results of a multi-path channel, as shown in Figure 1. This is due to wave reflections from hills, buildings, airplane flutter, and so forth. The effect of ghosts in an NTSC receiver can be very annoying. Nevertheless, an NTSC system transmits and receives analog video signals. Ghosts appear as multiple weakened and delayed images of the original picture. Although a picture could be severely ghosted, it is usually comprehensible to a viewer. This results in a "graceful" degradation, not a disaster.

On the other hand, for an all-digital HDTV system the video signal is heavily encoded as digital data. Reliable data recovery at the receiver is needed for the video decoder to store the original picture. However, when data is distorted by multi-path ghosts, even weak ones, the receiver might not be able to correctly decide which data was transmitted based on the received signal. This can severely damage the video decoding process. Therefore, if ghosts are not properly cancelled, broadcasting a heavily compressed digital HDTV signal via a multi-path channel can result in a problem much more serious than just ghosted pictures. In a severe case, the received picture could be totally incomprehensible to a viewer.

Most existing HDTV systems, such as the MUSE system (Multiple Sub-Nyquist Sampling) in Japan or the MAC system in Europe, use simple analog video encoding schemes based on sub-Nyquist rate sampling. The sampled ("discrete-in-time") analog video signal is pulse-shaped,[1] modulated and broadcast through DBS (Direct Broadcast Satellite). DBS provides a transmission channel which has a much less severe ghosting problem than broadcasting through a terrestrial channel. Also, due to the transmission of an analog encoded video signal, the effect of ghosts in DBS only results in very mildly ghosted pictures.

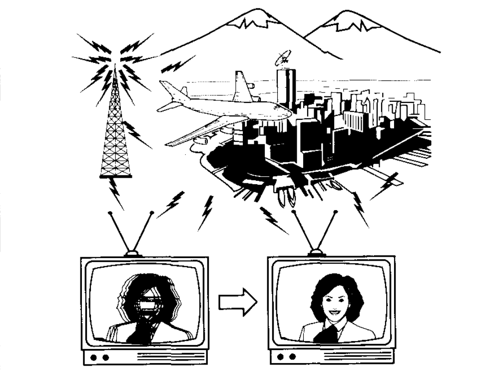

Figure 1: Effect of Ghost on NTSC Pictures

Other HDTV systems such as the ones proposed by Zenith and MIT for terrestrial broadcasting use a hybrid transmission scheme. The critical encoded video data and encoded audio data are sent through a highly protected data channel and other encoded video signals are sent as "discrete-in-time" analog signal components.

Before going further, we will describe the ghosting effect on a data transmission channel. The data transmission scheme that will most probably be used is the so-called Data Quadrature Amplitude Modulation (QAM), as opposed to the analog QAM with which TV engineers are very familiar. As shown in Figure 2(a), two-dimensional discrete digital data elements (a_n, b_n) are modulated and transmitted through an analog channel. As an

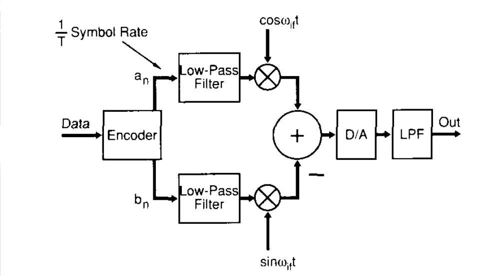

Figure 2a: QAM Transmitter Structure

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1 Raised-cosine pulse shaping, which satisfies the Nyquist criterion, is used to avoid intersymbol interference.

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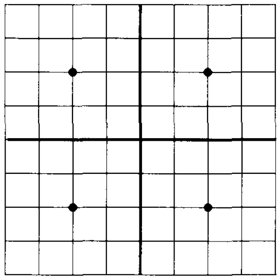

Figure 2b: 4-Point Signal Constellation

example, we only describe a simple system using four-point QAM. In this case, two consecutive data bits are mapped into a symbol which assumes one of the four points in a two-dimensional signal constellation, as shown in Figure 2(b). The symbol has two independent components called the inphase and quadrature components. They are further modulated with $\cos\omega_{\mathrm{if}}t$ and $-\sim \sin\omega_{\mathrm{if}}t$ respectively and summed before being up-converted for transmission. The IF spectrum is shown in Figure 2(c). When the signal is sent through the air, it is corrupted by multi-path distortion. After demodulation at the receiver, the received symbols are distorted and become very fuzzy as shown in Figure 2(d).

Note that the distorted signal around each symbol point in the constellation space resembles the original signal constellation with a reduction in strength. This is an interesting phenomenon caused by multi-path ghosts on data transmission. In Figure 2(d), we show a case with a mild multi-path distortion. If the multi-path distortion is severe enough, it might not be possible to correctly slice received symbols to the original transmitted symbols. This can be a disaster for a digital or hybrid HDTV receiver.

The greatest improvement of an enhanced NTSC system comes from the proper removal of ghosts from received pictures. Figure 3 shows a generic NTSC deghosting (multi-path equalization) block diagram which is described in a great detail in [1]. Multi-path channel equalization has also been a crucial issue for the success of a true HDTV broadcasting, either for DBS broadcasting or for terrestrial broadcasting which transmits either an encoded analog video signal or encoded digital video data. Figure 4 shows a data transmission system with a decision feedback multi-path equalizer at the receiver. However, equalizing such a channel at a very high digital signal processing rate over a wide ghost delay range is still very costly for current technology. This is particularly true for terrestrial broadcasting where ghosts are widely spread and an equalizer requires hundreds to thousands

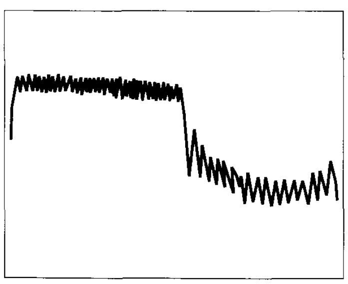

Figure 2c: Spectrum

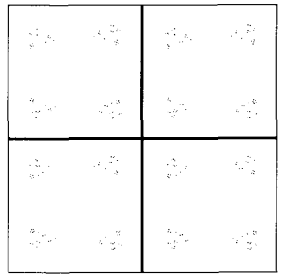

Figure 2d: Ghosted Signal Constellation

of filter taps at a rate around 15 M samples/second.

A companion paper [2] in preparation will deal with the implementation of several reduced-complexity equalizer structures for an enhanced NTSC receiver and for an HDTV receiver which utilizes digital communications. This paper will primarily deal with techniques for precise TV channel characterization. The information obtained from channel characterization can be used to set up a multi-path equalizer (ghost canceller). The channel information is valuable is many aspects which will be discussed in the next section.

II. CHANNEL CHARACTERIZATION FOR MULTIPATH EQUALIZATION

Channel characterization (identification) can be described as follows. At the transmitter, a known training signal (ghost cancellation reference signal, GCR) is sent through the unknown channel. At the receiver, the observed received signal is used to characterize the channel. We define $A(f)$ as the training (GCR) signal frequency spectrum, $H_t(f)$ as the transmitter frequency response, $H_c(f)$ as the multi-path frequency response, $H_u(f)$ as the tuner frequency response, $B(f)$ as the channel characterizer's frequency response at the receiver, and T(f) as the total frequency response.

$$\begin{aligned} T(f) &= A(f) \bullet H_t(f) \bullet H_c(f) \bullet H_u(f) \bullet B(f) \\ &= A(f) \bullet B(f) \bullet H(f), \quad H(f) = H_t(f) \bullet H_c(f) \bullet H_u(f), \end{aligned}$$

where $H(f)$ is the overall transfer function. If $A(f) \bullet B(f)$ equals a constant k over the transmission band, $T(f) \sim= k \bullet H(f)$, and the channel information is obtained. Going through a certain filter design procedure using channel information yields equalizer tap coefficients. This can speed up equalizer convergence in order to track a time-varying multi-path channel. Also, because the multi-

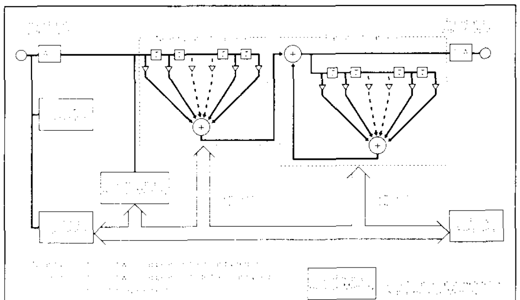

Figure 3: The General Structure of Ghost Reducer

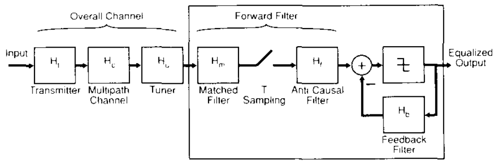

Figure 4: Multi-Path Equalization

path channel for TV broadcasting typically consists of relatively few significant and widely spaced responses, channel information can be used to constrain filter taps so as to reduce hardware complexity. Therefore, it would be useful to obtain precise channel information. In this paper, we will describe techniques for precisely characterizing a multi-path channel; one for an enhanced NTSC system and another for a spectrum compatible HDTV. Our proposed techniques can also provide various kinds of information for synchronization, automatic gain control, and carrier phase offset estimation. As an example, for an enhanced NTSC receiver our proposed GCR signal sequence can be used as a means to acquire synchronization, even if the received signal is severely ghosted or affected by impulse noise to the extent that the conventional synchronization scheme using synchronization pulses does not work.

If the channel is corrupted by noise, $T(f) = k \bullet H(f) + N(f)$, where $N(f)$ is the noise spectrum and k is the processing gain. The ratio $k/N(f)$, integrated over the transmission band, defines the signal-to-noise ratio. The larger the processing gain, the better the protection of channel information from the noise.

Review of the BTA's Approach

The BTA (Broadcasting Technology Association) in Japan proposed a simple channel characterization scheme for the enhanced NTSC system, which has a processing gain that is smaller than one. Conceptually, a truncated (in time) sin x/x pulse with a smooth passband of 4.2 MHz, surrounded by zero signal, is transmitted over a multi-path (ghosted) channel. At the receiver, a response composed of the sin x /x waveform and its associated ghosts is obtained. This response contains the multi-path channel information. The zero signal surrounding the sin x/x waveform is needed to warrant that the multi-path channel response is not interfered with ghosts of the preceding and the succeeding signals.

To avoid a possible DC bias problem in an NTSC system, the GCR signal is actually a truncated sin x/x pulse of 4.2 MHz bandwidth integrated over 44.7 μseconds and terminated by a 2T trailing edge (this signal is referred to as a wide-bar signal). This signal is inserted in line 18 during the blanking interval of a field and it is paired with a zero pedestal signal in line 18 of another field. A pair-wise fixed signal in line 17 at the above-mentioned fields is required in order to obtain interference-free GCR signals. After receiving the paired GCR signal, the ghosted zero pedestal signal in line 18 is subtracted from the ghosted integrated sin x/x GCR signal in line 18 of another field. By using this subtraction method, ghosts from the preceding pair-wise fixed signal are cancelled. This eliminates the need of sending surrounding zero signals. Therefore, using a limited time span a larger ghost

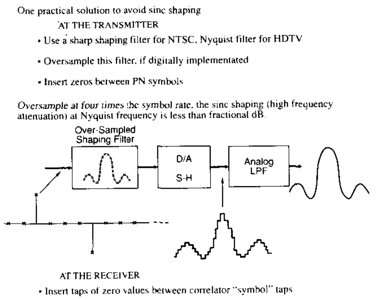

Figure 5: Pseudo Random Sequence Generating and Correlating Procedures

detection range can be accommodated. Another paired GCR signal is also transmitted at line 281 (253 lines or one field from line 18). To overcome the color burst phase reversal between fields, a paired GCR signal interleaved four times in an 8-field sequence format was designed by BTA [1]. The zero pedestal signal is sent in the 4^{th} field following its associated integrated sin x/x GCR signal. The paired GCR signal is interleaved four times in a 8-field sequence. (Other field sequences can also be used.) Conceptually, this GCR signal can be viewed as (sin x/x)/(1-D), where 1/(1-D) is a discrete integration operation at a sampling rate of four times color subcarrier, about 14.32 MHz. At the receiver, the integrated sin x/x signal goes through a (1-D) one-clock difference operation which undoes the integration operation at the transmitter, and also provides a DC blocking function in order to eliminate possible DC bias in the system.

At the transmitter, the peak amplitude of the integrated sin x/x GCR signal is limited to about 75 IRE to avoid a possible nonlinear distortion at the receiver under a severe ghosting condition. This limits the amplitude of the sin x/x signal before integration to about 38 IRE. Since the amplitude of the original sinx/x signal is about one half the amplitude of the integrated GCR signal, it results in a 6 dB loss in the signal-to-noise ratio. At the receiver, one-clock difference operation, which restores the sinx/x signal, results in a 3 dB noise enhancement, or equivalently a 3 dB reduction in the signal-to-noise ratio. The total loss due to the need of eliminating DC bias in the system is about 9 dB. A similar scheme was proposed for the MUSE HDTV system.[2] To improve the signal-to-noise ratio, the received GCR signal is averaged over several hundred fields before it is used. BTA proposed an average of about 5 seconds or 32 times over a 8-field sequence [1,4] to achieve 15 dB improvement in the signal-to-noise ratio. (Note that another 6 dB improvement in the signal-to-noise is achieved with an average over four pairs of the received GCR signal in a 8-field sequence.) The net improvement in 5 seconds is about 12 dB in the signal-to-noise ratio. In practice, when the channel is noisy, the 5-second averaging is performed over many iterations to obtain satisfactory channel information. This slows down the system convergence speed and makes the tracking of changing ghosts impossible.

In [3], results of field tests using the BTA approach was reported. This work was conducted by the National Association of Broadcasters (NAB) and the the Association of Maximum Service Television (MSTV) in Atlanta, Georgia, with supports from NHK (Nippon Hoso Kyokai), BTA, ATTC (Advanced Television Test Center), and many local stations. In that report, results showed

2 The MUSE system uses a sin x/x waveform which assumes a negligible DC bias in the system.

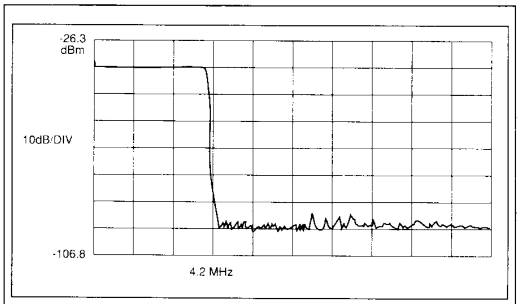

Figure 6a: The Power spectrum of PN Signal at the Correlator Output

that although the BTA system generally improved the picture quality, in some instances it actually reduced the quality of the received picture. Usually this condition occurred when the signal level was weak (low signal-to-noise ratio). This is worse than just slow down the system convergence. Note that the field tests were conducted within the City Grade services areas. It can be expected that this problem can be more severe in remote service areas.

As mentioned in the final report of BTA [4], the ghost strength and phase tend to change constantly. It is desirable that the cancelling time be held down to within 1 second, taking into account airplane flutter, compatibility with future sync broadcast services, and other factors. Apparently, the BTA's approach does not meet this requirement.

In [3], it was reported that homes with outdoor antennas displayed non-varying (stationary) ghosting conditions, and were largely corrected. However, homes with indoor antennas, such as rabbit-ears or monopoles for VHF and bow-ties or loops for UHF, experienced changing (dynamic) ghosts. These varying ghosting conditions were more prevalent where people were moving around in the room, swaying of trees, or vehicles passing by, etc. The BTA ghost canceller generally was not able to adequately compensate for, or track, these conditions. In these conditions, pseudo-ghosts (false ghosts) were actually added to an already ghosted picture. When a substantial (strong) ghost, usually a leading (pre-cursor) one, was encountered, the picture quality was also reduced. Malfunction of a synchronization separation circuitry under a strong ghosting condition might be attributed to this problem since the BTA system relies on an accurate count of horizontal synchronous pulses to determine the GCR signal insertion position. When this happens, the BTA system cannot acquire the GCR signal and therefore it does not work. In all, it was reported in [3] that the BTA system generally improved the picture quality by two steps based on a CCIR five-step ranking. However, the BTA system was unable to correct for ghosts in 7% of their observations. It can be expected that the percentage of failure of the BTA system should be higher if field tests were conducted in remote service areas. To partially solve problems encountered by using the BTA system, BTA [4] recommended to select the appropriate types of antenna, installation place, height, and direction. Apparently, this is not a desired solution in the United States of America.

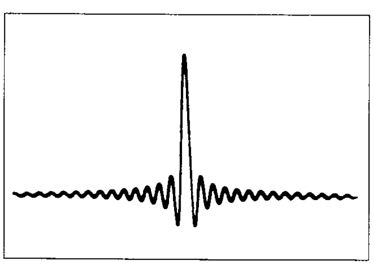

Figure 6b: The Time Domain Correlator Output of PN Approach

In the coming sections, we will propose a new approach that provides an opportunity to track changing ghosts at a weak signal level (low signal-to-noise ratio).

Modification of the PN Sequence and the Correlator Response

In our proposal to be described in this paper, a modified pseudo-random noise (PN) sequence and a properly designed correlator are used. Our correlator, matched to our modified PN sequence, provides 24 dB better noise rejection than that obtained from the use of a single pulse. No long-term averaging is needed to improve the signal-to-noise ratio. Therefore, it has a better system convergence speed and provides a hope for tracking changing ghosts, or it can trade-off the speed of convergence with the usage of VBI (vertical blanking interval) by sending the GCR signal less frequently. Our proposed correlator has a very low DC gain and provides an equivalent blocking function as the one-clock difference operation in BTA's proposal. Also, our proposed scheme provides rejection of other kinds of impairments, such as impulse noise, ignition noise, or colored noise, for which the long-term averaging is not effective.

One well-known method for channel identification is to send a pseudo-random noise sequence over a channel and to correlate the received sequence with that sequence. A unique property of a PN sequence with a length of N symbols is that when its repetitions are continuously correlated by the correlator, the correlator output results in large peaks separated by quiet zones of N-1 samples with a negative DC value, which is N times smaller than the large peaks in magnitude. When transmitting PN sequences over a channel with ghosting conditions, the quiet zone becomes corrupted with pulses which correspond to the ghosting conditions. In this quiet zone, each ghost appears as a weakened, delayed, and possibly phase-rotated replica of the direct signal. This permits the multi-path channel response to be characterized with respect to amplitude, phase, and delay. This method has the advantage of providing significant correlation gain against channel noise. However, there are many pitfalls in such an approach. One problem is that a quiet zone between the correlator peaks has nonzero DC values which would disturb the channel characterization of weak ghosts and possibly create problems for equalization. This problem can also be understood from a frequency domain point of view. The associated frequency spectrum of a nominal PN sequence is not flat. If N is the number of symbols in a PN sequence, the DC component has a magnitude that is $\sqrt{N}$ times in magnitude (or N times in power) smaller than the rest of the frequency components. For this reason, a nominal PN sequence is not adequate for exactly characterizing the channel information. Another problem is that correlating to the received input could require N multiplications-and-additions in each symbol period, which presents a significant overhead in hardware complexity. To overcome these problems, our schemes require modifications of the nominal PN sequence and its associated correlator.

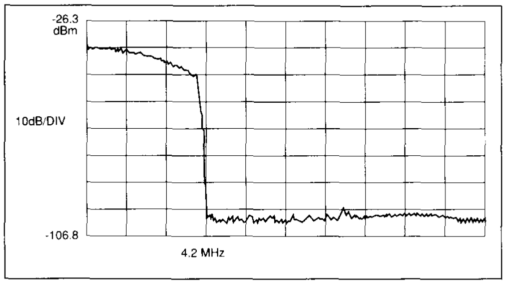

Figure 7a: The Power Spectrum of an Ill-Designed PN Signal at the Correlator Output

A PN sequence and the associated correlator will be modified in such a way that the product of their frequency spectra has a flat response in order to obtain an exact channel characterization. However, for simplicity in the circuitry, this flat response property should be obtained under the constraint that correlating with the received input only requires N additions or subtractions in each symbol period, with no multiplications. Therefore, the correlator can be cost-effectively implemented for a real-time operation. One can argue that this is not a concern if the correlation will be done on an off-line basis. However, our concern on the correlator implementation comes from three considerations: 1) future HDTV sets will most probably need to use the real-time correlator output as a means of synchronization, 2) future high-end NTSC sets might want to use the real-time correlator output for precise synchronization when the conventional synchronization mechanism fails under severely distorted conditions, 3) an NTSC ghost canceller without a new means of synchronization will fail under a severe ghosting condition where the conventional synchronization mechanism does not work. There are other constraints inherent to an NTSC system, although they might not be constraints for an HDTV system. One problem in a typical NTSC system is the DC bias problem in the demodulated signal. This can come from at least two sources: 1) a DC bias in a cost reduced synchronous detector of 3 to 4 IRE in magnitude. 2) the uncertainty in the pedestal region of ±2 IRE in magnitude. For this reason, another constraint in channel characterization, for an NTSC system and possibly for some HDTV system, is to provide a DC blocking effect at the correlator to significantly reduce the DC component. The channel information without DC bias can be used for calculating precise equalizer filter coefficients. Another problem of sending a PN sequence in an NTSC system is not to confuse a horizontal synchronization pulse or color burst with the strong negative values that a nominal PN sequence would have.

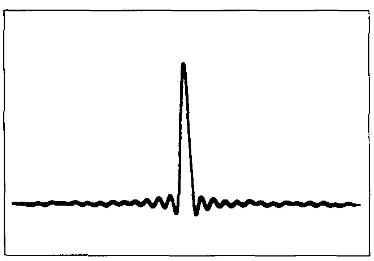

Figure 7b: The Time Domain Correlator Output of an Ill-Designed PN Approach

Before we discuss a "full-fledged" transmission system, for simplicity in the discussion we will only describe an equivalent discrete model which excludes the analog modulation and demodulation in the system. Later in this paper, we will described how to make the scheme work for two different transmission systems, the NTSC and the spectrum compatible HDTV. The reader is also referred to [5,6,7,8] for details.

There are at least three ways to modify the $A(f)$ and $B(f)$ spectra to make their product flat. (The combined spectrum of A(f) and B(f) will be referred to as the AB spectrum.) As mentioned above, the discrete spectrum of repeated PN sequences has a DC component which is $\sqrt{N}$ times in magnitude smaller than the rest of the frequency components. A simple modification is to add a small DC to a nominal PN sequence and amplify the DC component by a factor of $\sqrt{N}$. The bias constant can be found by the following formula

$$Bias = \frac{\sqrt{N+1}-1}{N}$$

Now, the GCR (training) frequency response is $A(f) = \sqrt{N}$. In this case, the channel characterizer is the correlator which mimics the modified PN sequence response. Since $B(f) = A(f)$ and $A(f) \bullet B(f) = N+1$, the frequency response at the correlator output is flat. For a modified PN sequence of a length of N = 255 symbols the correlation (processing) gain is about 24 dB better than the BTA's approach. When we consider the 3 dB noise enhancement by the (1-D) one-clock difference operation of the BTA's approach, the new scheme is 27 dB better. However, the bias constant is not an integer. The resulting correlator tap coefficients are non-integer.

AT TRANSMITTER	AT CORRELATOR
(1 + bias, -1 + bias)	(1 + bias, -1 + bias)

Correlating with the received GCR signal requires N multiplications-and-additions in each symbol period. Therefore, this approach does not satisfy the circuitry-simplicity constraint.

Another possibility of making the AB spectrum flat is for a correlator to mimic the nominal PN sequence response and to modify the PN sequence by amplifying its DC component by a factor of N. The overall DC amplitude becomes N, and the discrete AB spectrum is therefore flat. Following the above-mentioned approach yields a bias constant $\frac{\sqrt{N(N+2)+1}-1}{N}$. It is easy to verify that for any N this bias constant is 1.

The symbol value of this modified PN sequence has only one of the two values (2,0). These two values are normalized to (1,0) for simplicity in the discussion. The correlator tap coefficients have one of the two values (1,-1), where the locations of -1 tap coefficients correspond to the modified PN symbols of a zero value. Since the correlator tap coefficients are either 1 or -1, multiplications in correlating the received signal are not needed. Correlating with the input requires only N additions or subtractions in each symbol period. Note that this modified discrete PN sequence does not have negative values to confuse the horizontal synchronous pulse and the color burst.

AT TRANSMITTER	AT CORRELATOR
(1, 0)	(1, -1)

2 The MUSE system uses a sin x/x waveform which assumes a negligible DC bias in the system.

This approach satisfies all the constraints. It is easy to verify that $A(f) \cdot B(f) = (N+1)/2$. It provides about 21 dB correlation gain. It is 24 dB better than the BTA's approach.

Another possibility is to send the nominal PN sequence as the training signal and to have a correlator with a response having a DC component N times stronger than the rest of its frequency components. Now, the correlator tap coefficients have one of the two values (1,0) Correlating with the input requires N/2 additions in each symbol period.

AT TRANSMITTER	AT CORRELATOR
(1, -1)	(1, 0)

This approach requires the least correlator computation. However, this approach does not satisfy the constraint for an NTSC system where negative GCR values are not allowed and does not provide a DC blocking effect at the receiver. Nevertheless, this approach can be useful for an HDTV system if this constraint does not apply. In the rest of this paper, we will concentrate on the second approach which satisfies all the constraints.

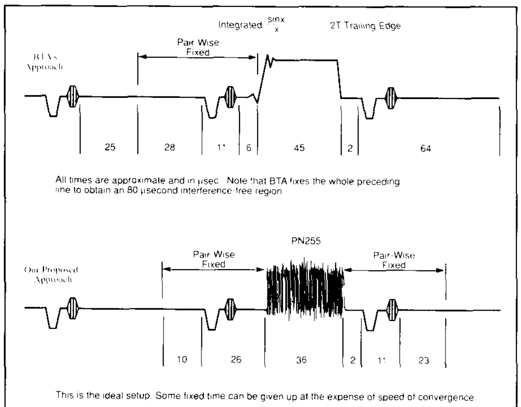

Figure 8: Comparison between BTA and Our Proposed NTSC GCR

Other well-known sequences such as a polyphase sequence [9,10] can also be used to obtain the correlation gain. A polyphase sequence has nice properties and it might seem to be more attractive than a PN sequence. However, for reasons to be discussed in Appendix A, we determined that a modified PN sequence was more suitable for our applications. Therefore, we will only deal with the modified PN sequence approach.

A property of this modified PN sequence is that when its repetitions are continuously correlated by the correlator,the correlator output results in large peaks separated by quiet zones of N-1 zero values. In the case of transmission over a ghosted channel, the signal corresponding to ghosts will appear in the quiet zones. This permits the multi-path channel response to be precisely characterized with respect to amplitude, phase, and delay. In the next section, we will show some examples.

Up to this point, we only considered the discrete PN sequence and correlator operation. In practice, the modified discrete PN sequence has to be properly shaped for transmission. This shaping should provide a flat frequency response over the desired transmission band. For an NTSC system, a low-pass filter with flat passband response and sharp attenuation at the cut-off frequency should be used. For an HDTV system, a raised-cosine pulse shaping filter should be used to avoid intersymbol interference between symbols. It is possible to do this in the analog domain. However, for accuracy we might want to implement filter shaping in the digital domain and then convert the digital signal to an analog signal for transmission. In this case, the sample-and-hold circuit in a digital-to-analog (D/A) conversion process introduces a step function between any two samples and results in a sin x/x shaping in the frequency domain. To overcome this problem, filter shaping should be operated at a rate several times higher than the symbol rate, and zero values should be inserted between any two discrete PN symbols at the input to the oversampled shaping filter, as shown in Figure 5. If the oversampling rate is four times higher than the symbol rate, the effect of the sin x/x frequency shaping on the transmitted signal is greatly reduced. The high frequency attenuation on the transmitted signal becomes negligible. At the receiver, for doing digital signal processing the received signal is usually oversampled. To avoid sin x/x high frequency attenuation at the correlator operation, the correlator tap coefficients should be properly spaced according to the oversampling rate. If the input signal to the correlator is sampled at twice the symbol rate, the correlator tap coefficients should be spaced by two samples. Conceptually, a null tap is inserted between every two active correlator taps. Figures 6(a) and (b) show the frequency-domain correlator output and the time-domain correlator output,

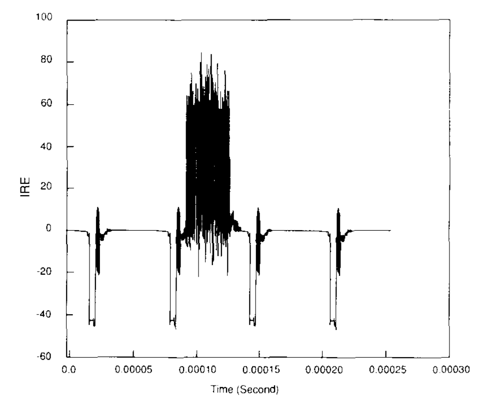

Figure 9: In-Phase Ghosted Training Signal

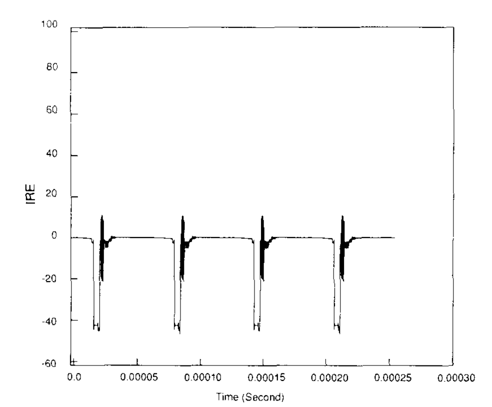

Figure 10: In-Phase Ghosted Blanking Signal

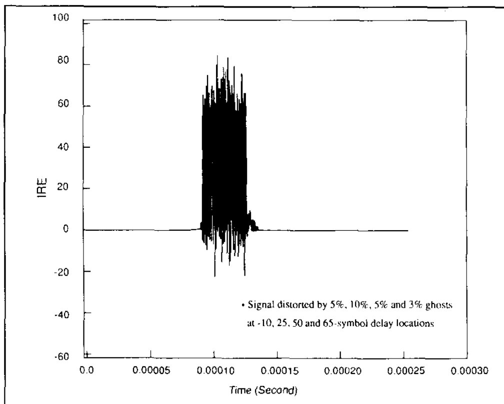

Figure 11: Remained Signal After Subtraction, Received by a Stable Pll TV Receiver

respectively. These can be compared to the results shown in Figures 7(a) and (b) of an ill-designed system where high frequency components are greatly reduced, such that the attenuation is about 13.4 dB at 4.2 MHz.

III. AN ENHANCED NTSC SYSTEM

If a sufficient time span was available, our proposed scheme would repeatedly send the shaped modified PN sequence at the transmitter. Therefore, at the receiver the received sequences could be correlated with its associated response to produce quiet zones for precise multi-path channel response characterization. However, a horizontal synchronization pulse is inserted between any two lines, and it is not possible to send many consecutive PN sequences without interruption. Also, the available bandwidth in the vertical blanking interval is precious. More and more demands for the usage of the VBI are foreseeable. To send repeated sequences is undesirable, if not impossible. As a result, our proposed scheme only involves the insertion of a single PN sequence of 255 symbols transmitted at 7.16 M symbols/second during the vertical blanking interval. In this case, a multi-path delay range of 35.6 μ second can be accommodated. At the receiver a new signal is created to mimic the same effect as if repeated sequences were sent at the transmitter. The choice of a symbol rate of 7.16 M symbols/second is due to a constraint in an NTSC receiver where the standard sampling rate is set at 14.32 MHz. To make it possible for us to design a correlator using only additions in the computation, we have to send a PN sequence at 7.16 M symbols/second which is an integral divisor of the sampling rate at the receiver. Then the tap coefficients can be set to 1 or -1. When the PN sequence at this symbol rate is shaped for transmission by a low-pass filter with a passband of 4.2 MHz, the overall correlation gain will be reduced by about 2.3 dB. Therefore, the net gain over the BTA's approach is about 21.7 dB.

According to a survey by BTA, 92% of ghosts are within a -4 to 26 μ second delay range and when extended to -4 to 37 μsecond, almost all cases are covered. This generally agrees with what is reported in [3] based on their limited observations in Atlanta, Georgia. While our approach can cover well beyond 92% of ghosts, it falls short by 5.4 μseconds for an almost perfect coverage. Nevertheless, even in the latter case, the picture quality is greatly improved since almost all ghosts are cancelled and only some extremely long ghosts remain on the picture. In Section 3.1, we will propose several ways to extend the ghost delay coverage

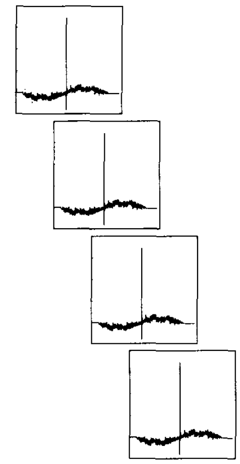

Figure 12a: Create Quiet Zones by Shift-and-Add

based on the same principle.

Similar to the BTA's approach, the transmitted sequence is preceded and succeeded by fixed signals (or at least pair-wise constant between successive repetitions) of duration at least that of the maximum expected multi-path delay. A matching line in a different vertical blanking interval contains a zero pedestal signal in place of the PN sequence, with the same leading and lagging fixed signals. As pointed out before, this can avoid interference from ghosts of the surrounding signals. If the modified PN sequence is placed right after the back porch (or about 2 μseconds after the color burst), about 23 μseconds of a pair-wise fixed signal should be used at the end of the preceding line and about 10 μseconds at the beginning of the succeeding line. If the modified PN sequence is placed right before the front porch (or about 2 μseconds before the horizontal synchronization pulse), about 10 μseconds of a pair-wise fixed signal should be used at the end of the preceding line and about 23 μseconds at the beginning of the succeeding line. Depending on how the VBI lines are used by broadcasters, one placement of the GCR signal could be more favorable than the other one. In Figure 8, we show a comparison between BTA's and our proposed NTSC GCR signals. In this figure, the modified PN sequence is placed right before the front

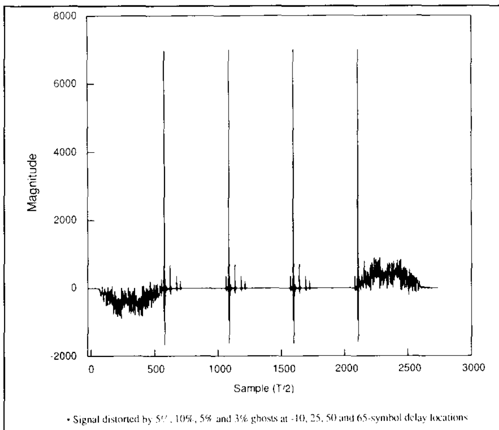

Figure 12b: In-Phase Correlator Output

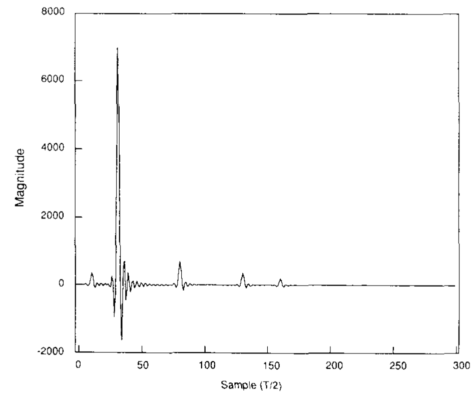

Figure 13: Characterized In-Phase Multi-Path Channel Response

porch. Note that BTA fixes the whole preceding line to obtain a 80 μ second interference-free region. (Ghosts with a delay longer than 45 μseconds and shorter than 80 μseconds do not interfere with ghost detection.) As to be discussed in Section 3.1, our scheme allows us to increase the size of fixed signals in order to significantly increase the ghost delay coverage.

To show another alternative, in the following example, we place the modified PN sequence right after the back porch. At the receiver, matching lines as shown in Figures 9 and 10 are subtracted, producing a modified PN sequence and its delayed versions due to multi-path, surrounded by zero signal, as shown in Figure 11. The signals are distorted by 5%, 10%, 5%, and 3% ghosts at -10, 25, 50, and 65-symbol delay locations. This new signal is then delayed three times by the length of the sequence, and the delayed sequences added to create four repetitions of the received PN sequence. The result is then fed to a correlator with +1 and -1 coefficients. The final correlator output contains four main peaks, at least one pair of which defines a quiet zone during which the multi-path responses can be clearly found. An equivalent approach, but computationally simpler, is to correlate the multi-path distorted PN sequence surrounded by zero signal and to delay the resulting signal at the correlator output three times by the length of the sequence, as shown in Figure 12(a); this new signal is added to create the final output, as shown in Figure 12(b). Figure 12(a) shows that the correlator output associated with a single ghosted modified PN sequence has a peak surrounded by some "sidelobe interferences" due to the transition periods of the correlator operation. Since any two samples of the correlator output separated by N symbols have the same magnitude and an opposite sign, the sidelobe interferences are cancelled after the "shift-and-add" operation. Therefore, the resulting output, as shown in Figure 12(b), has three quiet zones surrounded by sidelobe interferences. In theory, only the second quiet zone is truly quiet since the first and the last quiet zones are interfered by sidelobes of post-cursor ghosts and pre-cursor ghosts, respectively. Figure 13 shows a typical ghosted in-phase response in a single quiet zone. Figure 14 is the in-phase response for a perfect channel. The shift-and-add operation to mimic the transmission of consecutive PN sequences at the receiver would cause about 3 dB noise enhancement. However, when the signal in any two quiet zones are averaged the noise enhancement is about 1.5 dB. This is a penalty paid for our scheme. The net gain over the BTA's approach is about 20.2 dB. Note that when the discrete modified PN sequence is shaped by a low-pass filter with smaller bandwidth, intersymbol interference results. This causes overshoot and undershoot in magnitude. To avoid excessive overshoot and undershoot, the magnitude of the discrete modified PN symbol is reduced to 50 IRE to guarantee that the sending GCR signal does not exceed the transmitter's linear operation range and the ghosted GCR signal is well within the linear dynamic range of the receiver circuitry. This can be compared to the BTA's wide bar GCR signal which has a sin x/x waveform of about 38 IRE in magnitude before integration and 75 IRE at the leading edge after integration. This is about 2.4 dB in favor of the new approach. Therefore, the net gain of this approach is about 22.6 dB over the BTA's approach. To further warrant that the received ghosted GCR signal will not be distorted by nonlinearity and that the synchronization pulse separation circuit will not malfunction (accurate synchronization pulse separation before ghost cancelling is needed to determine the GCR signal insertion position), BTA [4] recommended to select the appropriate types of antenna, installation place, height, and direction so as to limit the ghost interference level to be 6 dB down from the direct signal. Further reduction in the GCR signal strength can avoid the possibility of nonlinear distortion in channel characterization. However, this will further reduce the signal-to-noise ratio. Since our scheme provides a significant correlation gain to begin with, we can afford to reduce the GCR signal strength. If the conventional synchronization pulse separation circuit malfunctions, the system fails. In this case, our approach provides a means to achieve precise synchronization using the peak in the real-time correlator output as a reference. If the simplified real-time correlator using only additions is still too complex, an even simpler correlator can be used. This correlator takes sliced signals as binary input and uses exclusive OR logic to match the PN sequence. The peak can than be used as a synchronization reference. The correlator used for channel characterization can be performed on an off-line basis to further reduce the cost.

Although it might not be needed, our scheme can be further improved by averaging over several 8-field sequences. In 0.25 second, which is still well within the 1.0 second specification of tracking changing ghosts [4], 9 dB improvement in the signal-to-noise can be achieved. Note that for the BTA's approach to obtain the same signal-to-noise ratio it takes about 68 seconds. For areas behind high building and hills the signal is weak. It has been observed that it can take longer than 68 seconds.

Confirming previous results by BTA [4], we also observed that a typical synchronous detector can be affected by phase noise [7].

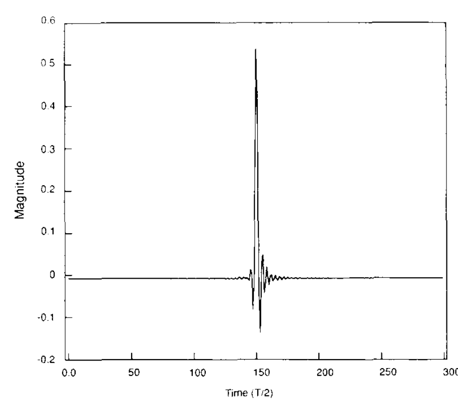

Figure 14: In-Phase Perfect Channel Response

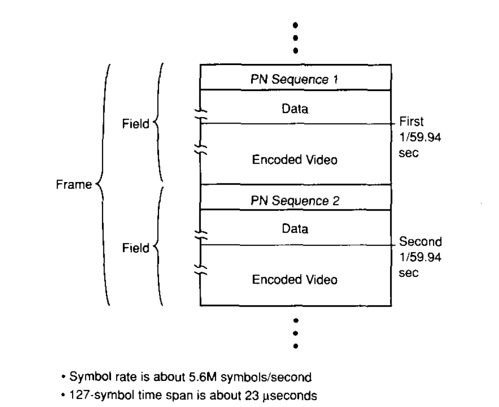

Figure 15: Spectrum Compatibility HDTV Frame Format

Therefore, the accuracy in channel characterization can deteriorate. However, our scheme provides a significant correlation gain so the effect is not severe. It appears as if there were some additional weak ghosts. Those false ghosts have a strength about 32 dB below the direct signal. A picture with those false ghosts will meet target values which are 30 dB below in a weighted measurement (or a subjective evaluation of score 4 based on a CCIR five-step ranking). Since score 4 is perceptible, but not annoying, our scheme allows a receiver to use a cheaper synchronous detector. Even if an ideal synchronous detector is used, the BTA's approach still has several drawbacks. For example, it provides a prolonged sync pulling duration at the time of power on or channel selection.

Note that for an NTSC system VSB modulation is used. When a ghost has a phase rotation with respect to the direct signal, the phase rotated quadrature component also appears in the inphase correlator output. This might seem to be an interference for channel characterization. In fact, this information is needed if we want to implement a "real" ghost canceller using only the inphase signal as the input, as opposed to a more costly cross-coupled (complex in mathematical sense) ghost canceller using both inphase and quadrature signals as the input. Therefore, only one correlator is used to acquire this complex channel information which includes the inphase channel response and the cross-coupled interference from the quadrature subchannel.

Extending Ghost Delay Coverage

As mentioned above, while our approach can cover well beyond 92% of ghosts, it falls short by 5.4 µseconds for an almost perfect coverage. In this section, several methods that can be used to extend the ghost delay coverage are discussed.

When a long delayed ghost appears in the channel, ambiguity in ghost detection results. For example, a post-cursor ghost delayed by y µseconds , which is longer than the maximum delay coverage, appears in the next quiet zone and is mistaken as a ghost delayed by (y - 35.6) µseconds. Also, a long ghost with a delay slightly shorter than the maximum delay coverage can be mistaken as a pre-cursor ghost. In this case, the multi-path equalizer would not cancel this long delay ghost but instead would create another false ghost. To prevent this from happening, we can change the symbol rate of the PN sequence to accommodate a larger ghost delay range. For example, a PN sequence of 255 symbols transmitted at 5.67 M symbols/second would accommodate a ghost delay range of more than 45 µseconds . Unfortunately, the correlator tap coefficients are not just 1 and -1. Correlating with the input at a rate of 14.32 MHz requires N multiplications and additions per symbol period to interpolate because of the non-integral ratio between symbol and sample rates.

In the above case, we assume that the "standard" sampling rate at the receiver (four times of color subcarrier frequency) should not be changed. If we are allowed to change the sampling rate at the receiver to a multiple of 5.67 MHz, such as 17.01 MHz, a simple correlator can still be used.

Note that the leading and trailing sidelobe interferences in Figure 12(b) are deterministic signals and thus, at least in theory, can be eliminated. Also, note that the leading period is not affected by the "shift-and-add" operation. Therefore, only pre-cursor ghosts exist in the leading period. Similarly, only post-cursor ghosts exist in the trailing period. Since almost all pre-cursor ghosts are very strong, most probably they can be detected even with the presence of the sidelobe interference. Even if this is not the case, with some elaboration, the sidelobe interference can be closely reproduced. As one example, the most damaging part of the sidelobe interference can be reproduced by convolving the channel response near the first peak with the known sidelobe interference of an ideal channel. Therefore, the leading sidelobe interference can be significantly reduced, so we can easily detect the locations of pre-cursor ghosts. The information can be used to resolve the ambiguity between a long delay post-cursor ghost and a pre-cursor ghost around the second peak in Figure 12(b). This allows channel characterization for pre-cursor ghosts with a delay up to -35.6 µseconds . Since post-cursor ghosts longer than the maximum delay coverage do not appear in the first quiet zone, the locations of post-cursor ghosts can be determined. The information can be used to identify any long delay ghost in the second quiet zone, which has an absolute delay between 35.6 µseconds and 71.2 µseconds . Note that we assume that ghosts delayed into the second quiet zone do not overlap with the short delay ghosts in their own correct locations and sufficient pair-wise fixed signals are placed before and after the GCR signals. In this case, we can identify ghosts with a delay between -35.6 µseconds and 71.2 µseconds.

Finally, note that if we borrow some ideas from a somewhat more complicated scheme used for SC-HDTV (to be described), the ghost delay range can be extended to -17.8 to 35.6 µ second. This requires alternatively sending two PN sequences whose length are 255 and 127, respectively.

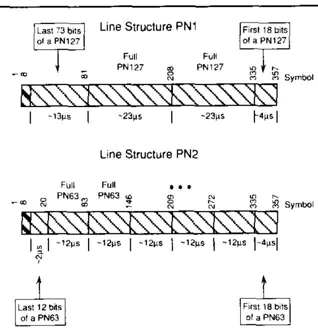

Figure 16

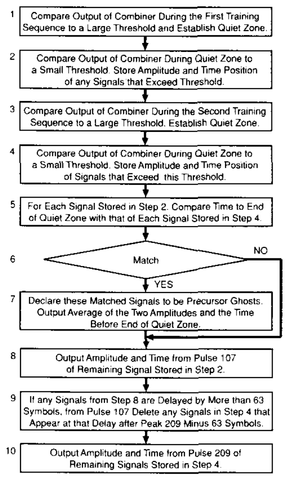

Figure 17: Ghost Ambiguity Resolution Algorithm

Trade-Off Between TV Ghost Cancelling Time And The Usage Of VBI

A concern of many broadcasters is the usage of the VBI resource. VBI is a finite resource and there is an increase in the demand for using VBI for many applications such as Teletext and data transmission. The usage of VBI to transmit the GCR signal is becoming a real concern. Since our new PN approach has a convergence speed that is several hundred times faster than BTA's wide bar GCR signal approach, our new PN approach offers a possibility of reducing the usage of VBI by slightly increasing the TV ghosting cancelling time. For example, the PN training signal can be sent one-tenth as frequently, thereby using one-tenth as much VBI resource, but still offers a speed of convergence that is an order of magnitude faster than BTA's wide bar GCR signal.

As proposed by BTA, in order to provide a wider ghost delay range with limited time span available to the GCR signal, "the subtraction method" is used. For BTA's approach a preceding line should only send pair-wise fixed signals. In our approach with 35.6 µsecond ghost delay coverage, the preceding line should have 10 µ second pair-wise fixed signal and 23 µseconds in the succeeding line. In North America, as authorized by the FCC, line 19 (field 1 and field 2) is dedicated to a fixed vertical interval reference (VIR) signal for automatic color adjustment at the receiver. Line 20 (field 1) is heavily used for channel ID and other information and field 2 carries a testing signal used by the networks. Line 21 (field 1 and field 2) sends captioning. Assuming that lines 19-21 cannot be changed for usage and line 18 will be used for broadcasting the GCR signal, BTA's approach requires that line 17 be used to send pair-wise fixed signals. Our approach only requires about 10 µseconds at the end of line 17 to send pair-wise fixed signals, and there are about 18 µseconds in line 18 that can also be used for sending pair-wise fixed signal. Therefore, the effective VBI usage of our approach is less than a line, not to mention the possibility of sending our proposed GCR signal less frequently to significantly reduce the VBI usage.

If the ghost delay distribution in a certain area is small (in most locations, it is very probable that ghost delay is less than 26 µseconds), an individual broadcaster can decide not to impose

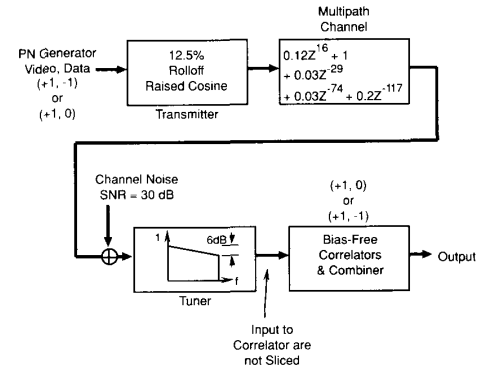

Figure 18:

any constraints on line 17, and our scheme would not be affected. Even if the ghost delay range is more than 36 µseconds and a broadcaster decides not to impose any constraints on line 17, our scheme will only be slightly affected. In this case, the long delayed ghosts from the last 10 µ second in the preceding line could corrupt, as noise, the near-by pre-cursor and post-cursor ghosts. Since long delayed ghosts are rare and weak and the near-by ghosts are strong the degradation might be negligible. Therefore, our scheme should perform well without enforcing any constraint on the use of line 17.

Finally, it is also possible to completely eliminate the need of sending pair-wise fixed signals. However, ghosts from the preceding line and signal in the succeeding line will corrupt, as noise, the channel characterization process. It will take many iterations of averaging before precise channel information can be obtained.

Comparison Between The BTA's Approach And Our Proposed NTSC Approach

Table 1 shows the comparison between the BTA's approach and our proposed NTSC GCR approach.

TABLE 1 Comparison

Metric	BTA's Proposal	Our Proposal
Against white noise	————	22.6 dB better
Ghost cancelling time	————	200 times faster
Usage of VBI	————	Can be reduced *
Tracking airplane flutter	Impossible	Possible
Loss of conventional synchronization before deghosting	Doesn't work	Still works
Degrading due to synchronous detector imperfection	Extreme	Slight
Degrading due to non-linearity	Extreme	Slight
Against interference other than white noise	Not effective	Effective
Overall complexity	Simple	Simple
Ghost delay coverage	44.7 µseconds	35.6 µseconds 106.8 µseconds **

* The GCR signal can be transmitted less frequently to reduce the usage of VBI at a small expense of system convergence.
* As mentioned in Section 3.1, with some elaboration it is possible for our scheme to extend the ghost delay coverage to -35.6 µseconds and 71.2 µseconds.

As reported in [3], the BTA system was unable to correct for ghosts in 7% of their observations, conducted in City Grade service areas. It can be expected that the percentage of failure of the BTA system is higher in remote service areas. Based on the

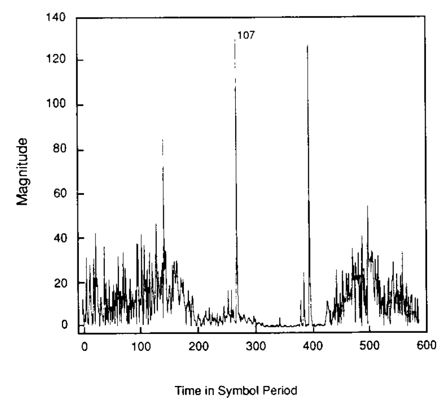

Figure 19: Correlator Output at 30 dB SNR with PN-127 Input

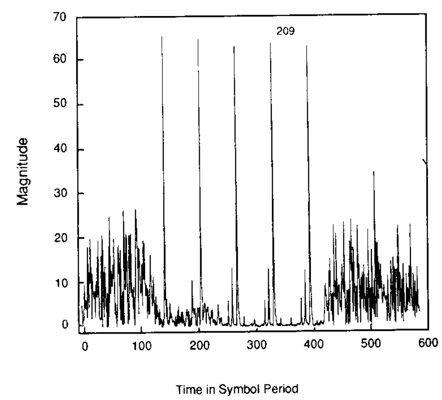

Figure 20: Correlator Output at 30 dB SNR with PN-63 Input

comparison made above, it is clear to us that our new approach can greatly improve the system performance and ghost correction coverage. The investments made by customers, broadcasters, chip suppliers, and TV manufacturers can therefore be protected.

IV. A TRAINING TECHNIQUE FOR HDTV

The same concept described above can also be used for HDTV applications. However, in most HDTV proposals, av.:ilable bandwidth is precious and the amount of time reserved for fixed signals must be kept to a minimum. In this section, we propose an alternative for HDTV applications, if this is desirable. The HDTV format used as an example here is shown in Figure 15.Only one line of 64 μsec duration is used for transmitting the training sequence during each field. Preceding and succeeding lines are used to carry other variable information, so the subtraction scheme outlined above cannot be used. Instead, two different repeated sequences are sent in alternate fields of each frame.

Note that an HDTV receiver using QAM modulation has independent inphase and quadrature components, and the equalizer should employ a cross-coupled (complex in mathematical sense) structure. To characterize this complex channel, a PN sequence is sent over either the inphase or quadrature subchannel. Some channel distortion can cause cross-interference between inphase and quadrature components. Therefore, at the receiver two correlators are used to acquire separate inphase and quadrature channel information for use in a complex equalizer. Note that this is different from an NTSC system using VSB where one correlator is used to acquire complex channel information, including the inphase response and the cross-coupled interference from the quadrature subchannel.

The example to be described is capable of characterizing multi-path responses with delays between -4 and 23 μsec., but can clearly be altered to handle other delay ranges. Both sequences are transmitted at a symbol rate of 5.6 MHz., +1 and -1 in level, shaped as before,and processed at the receiver with a correlator with coefficients of +1 and 0. Another choice is to transmit sequences of +1 and 0 in level, and processed at the receiver with a correlator with coefficients of +1 and -1. The former set of coefficients is illustrated in this paper.

The sequences are shown in Figure 16. The first sequence is of length 127, repeated twice with partial sequences before and after. The output of its correlator contains two main peaks, but the first 10 μ sec. of the interval may not be quiet because of delayed versions of preceding unrelated signals. The remainder of the interval is quiet, but signals appearing here may be either due to post-cursor multi-path delay between 13 and 23 μseconds , or to precursor multi-path.

The second sequence resolves these problems. This sequence is of length 63, repeated five times with additional partial sequences. The last pair of main pulses at the correlator output defines a quiet zone during which a different set of ambiguities exist. In that zone, ghosts due to a delay within 12 μseconds will overlap with post-cursor ghosts delayed from the sequence ahead and with precursor ghosts. However, by comparing the outputs of the two correlators, all ambiguities can be resolved as shown in the flow chart of Figure 17.

A simulated test was run using the channel of Figure 18, which includes noise and an imperfect tuner as well as multi-path. Figure 19 shows the output of the first correlator. The first part of the interval between main peaks is clearly corrupted, but multipath responses are readily extracted from the latter part of the interval, where the noise can be seen to be quite low. The output of the second correlator is shown in Figure 20, where the same multipath signals clearly appear. By comparing these two outputs, ambiguities can be resolved and the channel accurately characterized.

The use of PN training sequences is valuable for more than precise characterization of multi-path. The strong main pulse at the output of the correlator is an excellent signal to be used for synchronization and gain control.

APPENDIX A

V. CONSIDERATIONS OF USING A POLYPHASE SEQUENCE

Other well-known sequences such as a polyphase sequence [9,10] can also be used to obtain the correlation gain. A polyphase sequence has a frequency response with a equal magnitude values and many different phases which are roots of unity. It has zero autocorrelation between peaks. It has been shown that sequences exist for all lengths. If the length N is odd, an N-phase sequence of real values can be constructed; if N is even, 2N phases are needed. For convenience in the discussion, we will refer to these sequences as real polyphase sequences. With a minor extension, we can construct a complex polyphase sequence composed of an inphase sequence and a quadrature sequence. Those two sequences are orthogonal with zero crosscorrelation. In this case the sequence length is doubled. The frequency response of the inphase sequence is constructed by inserting a zero between any two samples of the real polyphase sequence. The frequency response of the quadrature sequence can be constructed in a similar way. The zero values of the two orthogonal sequences are offset by one sample to warrant that the product of the two spectra is zero (disjoint) and therefore, orthogonal. From a time domain point of view, a complex polyphase sequence of length N has an inphase correlation function with two positive peaks of the same magnitude at the $(N/2)^{th}$ sample and the N^{th} sample, respectively, and a quadrature correlation function with two peaks of different sign and the same magnitude at the $(N/2)^{th}$ sample and the N^{th} sample, respectively. The other correlation values are zeros. Note that the ghost delay coverage of a complex poly-phase sequence is only half of its length.

For an NTSC system using VSB modulation, a real polyphase sequence can be used. For an HDTV system using QAM with independent inphase and quadrature components (QAM is conceptually a complex channel), a complex polyphase sequence of orthogonal components can be used. (We will describe in Section 4 how a real sequence can be used to characterize a complex channel.) A polyphase sequence has nice properties and it might seem to be more attractive than a PN sequence. For example, the length of a real polyphase sequence can be any positive integer, as opposed to a PN sequence where the length needs to be $2^n - 1$, n a positive integer. The frequency response of a polyphase sequence is flat, as opposed to a PN sequence which has a small DC component. However, correlating with a polyphase sequence requires non-integral tap coefficients. Therefore, it always requires N multiplications and additions in each symbol period. A polyphase sequence has significant negative magnitudes which might interfere with NTSC synchronization pulses and color burst in an existing TV receiver. The correlator associated with a polyphase sequence does not provide a DC blocking effect. For a complex polyphase sequence the ghost delay coverage is only one half of its length. Therefore, it results in degradation for some equalization schemes. Due to the considerations mentioned above, we only deal with the modified PN sequence approach in this paper.

APPENDIX B

VI. THE EFFECT OF HIGH FREQUENCY ATTENUATION IN GHOST CANCELLATION

Reference [1] describes a "forward algorithm"[3] which takes the fast Fourier transform (FFT) and inverse FFT of the received reference signal and a "dividing method" to obtain the equalizer tap coefficients. If high frequency attenuation occurs, the tap coefficients are not accurate and the equalizer can't recover the higher frequency components such as the color signal.

Another method is the "feedback algorithm"[4] also described in [1]. The channel is characterized using the received reference signal in order to obtain the information about where ghosts are located. The LMS (Least-Mean-Squared) adaptive algorithm, which is constrained in the number of tap locations (sparse filter), is then used to fine tune the tap coefficients, with the received reference signal as the input and the original reference signal as the ideal signal. An additional effect of high frequency attenuation is to slow down the speed of convergence of the adaptive LMS algorithm due to the spreading in eigenvalues associated with the input.

It might be worthy of mentioning that the feedback algorithm is much more sensitive to DC offset in the system than the forward algorithm.

There are other filter tap coefficient calculation algorithms [2] which are more suitable for VLSI implementation. They should be used in order to implement a cost-effective ghost cancelling VLSI circuitry.

APPENDIX C

VII. USE OF PN TRAINING SIGNAL FOR OTHER SIGNAL PROCESSING

Co-channel interference will be a serious problem for simulcasting HDTV and NTSC signals. For example, an HDTV system in this environment experiences strong narrow-band carrier interference from a NTSC channel. The PN approach can provide correlation gain to minimize the effect of this narrow-band interference. This provides several hundred times better narrow-band interference immunity over BTA's wide bar GCR signal. Ideally, for an HDTV system a fixed comb filter can be used to reduce the NTSC co-channelnarrow-band interference. The picture carrier in an NTSC signal can drift by a FCC regulated nominal value, ± 1 kHz.[5] If the correlation gain does not provide enough immunity for very accurate channel characterization, some type of adaptive noise cancellation technique should be used. An adaptive harmonic (comb) noise canceller [11] can be used to mitigate this problem. This adaptive noise canceller can position itself according to the frequency drift in the carrier. However, to properly adapt to a harmonic noise canceller, the main signal should be highly distinguishable in correlation from the narrow-band interference. The PN signal is an excellent candidate for this type of signal processing but BTA's wide bar GCR signal is not. Our hardware implementation using this adaptive harmonic noise canceller working in conjunction with the correlator shows that when the PN reference signal is corrupted by a narrow-band interference our scheme creates a harmonic notch filter response in the frequency spectrum to cancel this interference. Therefore, it improves the overall performance. There are many more possible signal processing techniques that are applicable to improve the picture quality, and most of them can take advantage of a training signal of the desired correlation property of the PN signal.

3 The forward algorithm does not imply the use of a FIR filter in the cancellor. In fact, [1] describes a FIR (forward) filter to cancel near-by ghosts and a IIR (feedback) filter to cancel far ghosts.

4 Similarly, the feedback algorithm does not imply the use of an IIR filter in the canceller.

5 Today's carriers have a much more stable carrier. The offset is about 100 Hz. Technology is available to have an even more precise carrier with an offset of 1 Hz.

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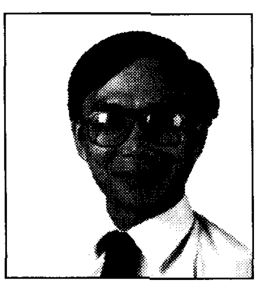
BIOGRAPHY

Jin-Der Wang was born in Chia-Yi, Taiwan, on November 13, 1955. He received the B.S.E.E. and M.S.E.E. degrees from the National Chiao-Tung University, Taiwan, in 1978 and 1980, respectively, and Ph.D. degree in Electrical and Computer Engineering from North Carolina State University, Raleigh, in 1985.



During his military service, form 1980 to 1982, he was an Instructor in the Department of Physics and Electrical Engineering, Naval Academy, Taiwan.
He joined the Data Communications Research Department at AT&T Bell Laboratories, Middletown, NJ, in 1985, as a member of the Technical Staff. His current research interests are in Communication theory, high-speed adaptive signal processing, television signal processing, and analysis of various data networks.

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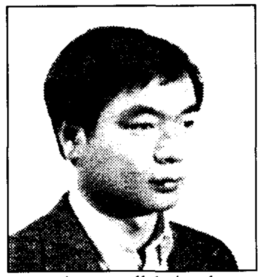

From 1985 to 1988, he was with Systems Technology Research Group, Television Research Lab., David Sarnoff Research Center, Princeton, NJ. Since 1989, he has been an associated professor in the National Chiao-Tung University, Taiwan.

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