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(Working Draft) Informational Document: DTV Signal Reception and Processing Considerations

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The Advanced Television Systems Committee, Inc., is an international, non-profit organization developing voluntary standards for digital television. The ATSC member organizations represent the broadcast, broadcast equipment, motion picture, consumer electronics, computer, cable, satellite, and semiconductor industries. Specifically, ATSC is working to coordinate television standards among different communications media focusing on digital television, interactive systems, and broadband multimedia communications. ATSC is also developing digital television implementation strategies and presenting educational seminars on the ATSC standards.

ATSC was formed in 1982 by the member organizations of the Joint Committee on InterSociety Coordination (JCIC): the Electronic Industries Association (EIA), the Institute of Electrical and Electronic Engineers (IEEE), the National Association of Broadcasters (NAB), the National Cable Television Association (NCTA), and the Society of Motion Picture and Television Engineers (SMPTE). Currently, there are approximately 160 members representing the broadcast, broadcast equipment, motion picture, consumer electronics, computer, cable, satellite, and semiconductor industries.

ATSC Digital TV Standards include digital high definition television (HDTV), standard definition television (SDTV), data broadcasting, multichannel surround-sound audio, and satellite direct-to-home broadcasting.

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1. SCOPE

This report describes various aspects of VSB DTV signal reception and processing. This document is intended to compliment ATSC Recommended Practice A/54A, "The Guide to the Digital Television Standard" [3], which provides an overview of DTV receivers in Section 9. This report explores many aspects of receiver design in greater detail, with considerable emphasis on equalization techniques. The following sections provide references to literature descriptive of VSB DTV receiver design.[1]

2. REFERENCES

2.1 Normative References

There are no normative references.

2.2 Informative References

1. ATSC Standard A/52A (2001): "Digital Audio Compression (AC-3)," Advanced Television Systems Committee, Washington, D.C., August 20, 2001.
2. ATSC Standard A/53B (2001): "ATSC Digital Television Standard," Advanced Television Systems Committee, Washington, D.C., August 7, 2001.
3. ATSC Recommended Practice A/54A (2003): "Guide to the Digital Television Standard," Advanced Television Systems Committee, Washington, D.C., 2003.
4. ISO/IEC IS 13818-1:2000 (E), International Standard, Information technology – Generic coding of moving pictures and associated audio information: Systems.
5. ISO/IEC IS 13818-2, International Standard (1996), MPEG-2 Video.

3. DEFINITIONS

With respect to definition of terms, abbreviations, and units, the practice of the Institute of Electrical and Electronics Engineers (IEEE) as outlined in the Institute's published standards shall be used. Where an abbreviation is not covered by IEEE practice, or industry practice differs from IEEE practice, then the abbreviation in question will be described in Section 3.1 of this document.

3.1 Terms Employed

For the purposes of the Digital Television Standard, the following definitions apply:

ACATS Advisory Committee on Advanced Television Service.

A/D Analog to digital converter.

anchor frame A video frame that is used for prediction. I-frames and P-frames are generally used as anchor frames, but B-frames are never anchor frames.

ANSI American National Standards Institute.

[1] Much of this literature is found in U. S. patents. Patent references in this document 9 should be regarded solely as technical literature. ATSC makes no representations concerning the validity or scope of the patents, or efficacy of the technology.

ATTC Advanced Technology Test Center.

AWGN Additive white Gaussian noise.

bidirectional pictures or **B-pictures** or **B-frames** Pictures that use both future and past pictures as a reference. This technique is termed *bidirectional prediction*. B-pictures provide the most compression. B-pictures do not propagate coding errors as they are never used as a reference.

bit rate The rate at which the compressed bit stream is delivered from the channel to the input of a decoder.

block A block is an 8-by-8 array of pel values or DCT coefficients representing luminance or chrominance information.

bps Bits per second.

byte-aligned A bit in a coded bit stream is byte-aligned if its position is a multiple of 8-bits from the first bit in the stream.

CDTV See *conventional definition television*.

channel A digital medium that stores or transports a digital television stream.

CIR Channel impulse response.

coded representation A data element as represented in its encoded form.

compression Reduction in the number of bits used to represent an item of data.

constant bit rate Operation where the bit rate is constant from start to finish of the compressed bit stream.

conventional definition television (CDTV) This term is used to signify the *analog* NTSC television system as defined in ITU-R Recommendation 470. See also *standard definition television* and ITU-R Recommendation 1125.

CRC The cyclic redundancy check used to verify the correctness of the data.

data element An item of data as represented before encoding and after decoding.

decoded stream The decoded reconstruction of a compressed bit stream.

decoder An embodiment of a decoding process.

decoding (process) The process defined in the Digital Television Standard that reads an input coded bit stream and outputs decoded pictures or audio samples.

discrete cosine transform A mathematical transform that can be perfectly undone and which is useful in image compression.

DTV Digital television, the system described in the ATSC Digital Television Standard.

D/U Desired (signal) to undesired (signal) ratio.

elementary stream (ES) A generic term for one of the coded video, coded audio, or other coded bit streams. One elementary stream is carried in a sequence of PES packets with one and only one stream_id.

encoder An embodiment of an encoding process.

encoding (process) A process that reads a stream of input pictures or audio samples and produces a valid coded bit stream as defined in the Digital Television Standard.

field For an interlaced video signal, a "field" is the assembly of alternate lines of a frame. Therefore, an interlaced frame is composed of two fields, a top field and a bottom field.

FIR Finite-impulse-response.

FPLL Frequency and phase locked loop.

frame A frame contains lines of spatial information of a video signal. For progressive video, these lines contain samples starting from one time instant and continuing through successive lines to the bottom of the frame. For interlaced video, a frame consists of two fields, a top field and a bottom field. One of these fields will commence one field later than the other.

group of pictures (GOP) A group of pictures consists of one or more pictures in sequence.

high-definition television (HDTV) High-definition television is a system which provides significantly improved picture quality relative to conventional (analog NTSC) television, with more visible detail, a wide screen format (16:9 aspect ratio), and may be accompanied by digital surround-sound capability.

Huffman coding A type of source coding that uses codes of different lengths to represent symbols which have unequal likelihood of occurrence.

IEC International Electrotechnical Commission.

IIR Infinite-impulse-response.

ISO International Organization for Standardization.

ITU International Telecommunication Union.

JEC Joint Engineering Committee of EIA and NCTA.

LMS Least mean squares.

Mbps 1,000,000 bits per second.

MPEG Refers to standards developed by the ISO/IEC JTC1/SC29 WG11, *Moving Picture Experts Group.* MPEG may also refer to the Group itself.

MPEG-2 Refers to ISO/IEC standards 13818-1 (Systems), 13818-2 (Video), 13818-3 (Audio), 13818-4 (Compliance).

picture Source, coded, or reconstructed image data. A source or reconstructed picture consists of three rectangular matrices representing the luminance and two chrominance signals.

pixel "Picture element" or "pel." A pixel is a digital sample of the color intensity values of a picture at a single point.

pre-echo An RF signal echo received via a path that is shorter than the path over which the principal signal is received.

program A program is a collection of program elements. Program elements may be elementary streams. Program elements need not have any defined time base; those that do have a common time base and are intended for synchronized presentation.

receiver A receiving device intended to tune and decode ATSC-compliant digital television signals.

ROM Read-only memory.

SAW filter Surface-acoustic-wave filter.

SMPTE Society of Motion Picture and Television Engineers.

splicing The concatenation performed on the system level of two different elementary streams. It is understood that the resulting stream must conform totally to the Digital Television Standard.

standard definition television (SDTV) This term is used to signify a *digital* television system in which the quality is approximately equivalent to that of NTSC. This equivalent quality may be achieved from pictures sourced at the 4:2:2 level of ITU-R Recommendation 601 and subjected to processing as part of bit rate compression. The results should be such that when judged across a representative sample of program material, subjective equivalence with NTSC is achieved. See also *conventional definition television* and ITU-R Recommendation 1125.

TOV Threshold of visibility, defined as 2.5 data segment errors per second.

trellis code A code generated by a finite-state machine, in which input of the information to be coded causes transitions or branches among states, each branch accompanied by the output of a group of encoded symbols.

variable bit rate Operation where the bit rate varies with time during the decoding of a compressed bit stream.

code in the data stream.

8 VSB Vestigial sideband modulation with 8 discrete amplitude levels.

16 VSB Vestigial sideband modulation with 16 discrete amplitude levels.

4. OVERVIEW OF THE DTV RECEIVING SYSTEM

A receiving device intended to tune and decode ATSC-compliant digital television signals (the *receiver*) must perform the following basic functions:

- Tune the selected 6 MHz channel
- Reject adjacent channels and other sources of interference
- Perform needed synchronization to identify symbols, data segments, training signals, etc.
- Apply required gain adjustment (AGC) to maintain the optimum noise figure
- Demodulate (equalize as necessary) the received signal, applying error correction to produce a transport bit stream
- Identify the elements of the bit stream using a transport layer processor
- Select each desired element and send it to its appropriate processor
- Decode and synchronize each element
- Present the programming

Specific issues affecting receiver design are discussed thoroughly in the following sections. In general, this document focuses on recovery and demodulation of the terrestrial-broadcast RF signal, because it is the most challenging of receiver processes. Noise, interference, and multipath are elements of the terrestrial transmission path, and the receiver circuits are expected to deal with these impairments. Innovations in equalization, automatic gain control, interference cancellation, and carrier and timing recovery (synchronization) create product performance differentiation and improve signal coverage.

The decoding of transport elements that make up the programming is usually considered to be a more straightforward implementation of specifications, although opportunities for innovation in circuit efficiency or power usage exist. In particular, innovations in video decoding offer opportunities for savings in memory, circuit speed, and complexity. The user interface and new data-based services are important areas of product differentiation.

For reference, a block diagram of the Grand Alliance prototype receiver is shown in Figure 4.1. A detailed discussion of the elements of this design can be found in [3].

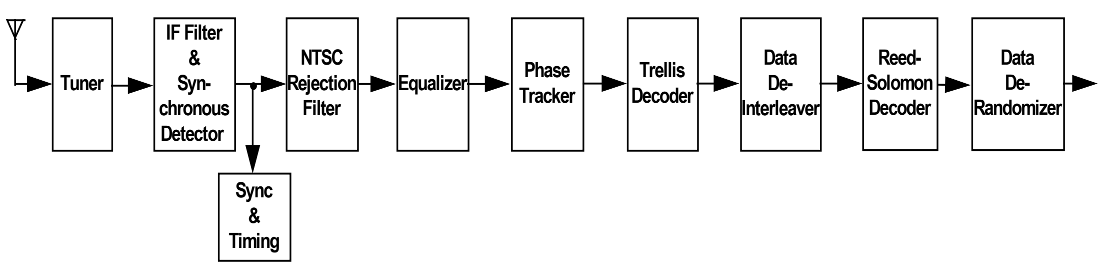

Figure 4.1 Block diagram of Grand Alliance prototype VSB receiver.

5. RECEIVER EQUALIZATION ISSUES

The multipath distortion of a received DTV signal comprises an ensemble of "echoes" or "ghosts" that accompany the principal component of the received DTV signal because of the signal being received via various paths. Echoes received via paths that are shorter than that over which the principal signal is received are termed "pre-echoes". Echoes received via paths that are longer than that over which the principal signal is received are termed "post-echoes". The DTV receiver contains channel-equalization and echo-suppression filtering for selecting the principal signal by suppressing any accompanying echoes that have sufficient strength to cause errors during data-slicing. This filtering also corrects in some degree for the receiver not having optimal frequency response; e. g., owing to tilt in the IF amplifier response.

The ATSC Digital Television Standard [2] specifies no particular design for the channel-equalization and echo-suppression filtering in the DTV receiver. When designing an adaptive equalizer, usually a designer considers and evaluates various possible ways of implementing this filtering. Section 5.4 treats general issues concerning the choice of equalizer implementation and discloses the particular combination of techniques used in the Grand Alliance prototype receiver [3]. Specifics concerning alternative designs for the adaptive equalizer in the DTV receiver will be found in Section 6.13, *infra*.

5.1 Echo-Suppression Range

Echoes are considered to be of substantial strength if their presence causes a readily observable increase in the number of data-slicing errors. There have been reports of pre-echoes of substantial strength being observed in the field that are advanced as much as 30 microseconds respective to the principal signal. There have also been reports of post-echoes of substantial strength being observed in the field that are delayed as much as 60 microseconds respective to the principal signal. The occurrence of such far-advanced or long-delayed echoes are peculiar to a limited number of reception sites. Most DTV receivers have equalizers with substantially less than 90-microsecond echo-suppression range, in order that the utility/cost ratio is better for the public in general.

Equalizers capable of suppressing pre-echoes advanced no more than 3 microseconds and delayed no more than about 45 microseconds were the general rule in the 1995 to 2000 timeframe. Since 2000 there has been increased awareness that channel equalizers should have capability for suppressing further-advanced pre-echoes. Also, with possibility of various field implementations of Distributed Transmission Systems, the need for longer pre-echo handling might be necessary.

5.2 Bit-Resolution of Baseband DTV Signal and Precision of Weighting Coefficients in the Equalization Filter

Increase in the bit-resolution of the DTV signal supplied for equalization increases the cost of the A/D conversion apparatus. The cost of the equalizer filter also increases as the bit-resolution of that DTV signal increases. Each weighting coefficient of the equalizer should have sufficient resolution that the product of that coefficient and the DTV signal has several more bits resolution than the DTV signal, so the weighted sum resulting from adding those products maintains the resolution of the DTV signal after it has been equalized. If there is insufficient resolution in the products of the weighting coefficients and the DTV signal, using these products to suppress echoes will introduce sufficient quantization noise into the equalizer filter response to increase data-slicing error appreciably, especially when there are strong echoes.

At least 10 bits resolution of the DTV signal supplied for equalization is generally regarded to be necessary if the performance of an 8-VSB receiver is to be commercially acceptable.

Increasing the DTV signal resolution a few more bits will facilitate equalization under more difficult multipath reception conditions, provided received signal strength is substantially larger than Johnson noise. This is because quantization noise (introduced by the equalizer) will not reduce SNR as much, and because weaker-strength frequency components of the signal are less apt to be irretrievably lost in quantization.

If fixed-binary-point multiplication is used in a time-domain equalization filter, weighting coefficients with at least 11 bits resolution are used. The specification of 11 bits resolution, minimum, is based on the guess that there are usually no more than 128 taps in the equalizer filter that have non-zero weighting coefficients. Increasing the resolution of the weighting coefficients a few more bits will accommodate more taps in the equalizer filter having non-zero weighting coefficients, a condition likely to occur under very difficult multipath reception conditions. By 2002 the state of the art in monolithic integrated circuitry had advanced sufficiently that the earlier constraints on equalizer design were considerably eased and weighting coefficients with 13 bits resolution were the commercial norm. In equalizers that use floating-point multiplication, the resolutions of the weighting coefficients can be adjusted to values needed for maintaining the resolution of the DTV signal after it has been equalized.

The resolution of the DTV signal after it has been equalized is kept higher than the four-bit minimum required for data slicing, so there is the extra resolution needed to implement Viterbi decision making.

5.3 Background Issues Concerning Equalization

There are some issues concerning choice of equalizer design that have been resolved sufficiently well that they are no longer considered foreground issues.

Most designers prefer equalization of single-carrier digital modulation to be done at baseband rather than in the IF passband. One reason for this is that equalization of real-only signals can be used, avoiding equalization of complex signals. Another reason is that the latent delay of re-modulation can be avoided in decision feedback. (See Section 6.13.6 “Passband Equalization” for more information concerning equalization in the IF passband.)

Most designers prefer that equalization be done in the time domain rather than in the frequency domain. Frequency-domain equalization can be used to boost energy in portions of the frequency spectrum where nulls occur. However, noise energy as well as signal energy is boosted in those portions of the frequency spectrum. This boost in noise energy is a substantial reason for performing time-domain equalization instead. (See Section 6.13.7 “Frequency-Domain Equalization” for more information concerning equalization being performed in the frequency domain.)

5.4 Equalizer Filter Implementation in the Time Domain

Most if not all receiver designs commercialized in 2002 or before used time-domain channel-equalization and echo-suppression filtering, operative on digitized baseband signals. The equalizer in the Grand Alliance prototype receiver is a transversal filter for digitized baseband signal, the weighting coefficients of which are adjusted by adaptive filtering procedures. The equalizer cascades an infinite-impulse-response (IIR) filter after a finite-impulse-response (FIR) filter operated as a feed-forward filter and employs decision feedback.

Some receiver designers prefer an equalizer design that uses just an FIR filter operated as a feed-forward element and does not include an IIR filter. However, many, if not most, designs for consumer-market receivers include an IIR filter. The IIR filter may be constructed so as to employ a feedback FIR filter within a recursion loop.

In the design for a time-domain equalizer, a feed-forward FIR filter has the attractive properties of always being stable and being able to cancel both pre-echoes and post-echoes. Unfortunately, in the process of canceling a primary echo, a feed-forward FIR filter generates a first-repeat (pre- or post-) echo of opposite polarity from the primary echo. This first-repeat echo has twice the delay (or advance) with respect to the primary echo and has a normalized amplitude that is the square of the primary echo (normalization being such that the principal signal has unit amplitude). This first repeat echo is then canceled in a similar manner, with a second-repeat echo being generated opposite in polarity from the first-repeat echo and with normalized amplitude that is the square of the first-repeat echo. Provided the kernel of the FIR filter provides a sufficiently large window in time, this process continues until the n^{th}-repeat echo generated is so small as to be lost in the noise of the signal, or is lost in the quantization process of the digital filtering. (This amplitude is found to be 20dB below the primary signal.) When an original echo has larger energy respective to the principal (or "cursor") component of the signal supplied for equalization, the repeat echoes with significant energy spread over a greater time. Therefore, in order for a feed-forward FIR filter to be able to cancel long-delayed post-echoes that have substantial energy as compared to the principal component of the signal supplied for equalization, the kernel of that FIR filter must span a great number of symbol epochs. Fortunately, there is a tendency for post-echoes that are longer delayed to be weaker than less-delayed components of the signal, reducing the likelihood of their repeats having substantial energy.

There tends to be substantial growth of noise level, called Equalizer Noise, during the echo suppression process in a feed-forward FIR filter. This process involves combining differentially delayed, weighted responses to the input signal of the feed-forward FIR filter to generate its response and suppress the echoes in that response. Some of the noise growth can be attributable to the combined quantization noise of the many weighted terms that are required when there is a number of echoes with substantial enough energies to require their suppression. Noise growth arising in the equalizer from combined quantization noise can be kept small by choosing sufficiently large bit resolution of the signal supplied for equalization and sufficiently large bit resolution of the weighting coefficients used in transversal filtering. Some of the noise growth in the feed-forward FIR filtering process can be attributable to the combined (digitized) Johnson noise of the many weighted terms that are required when there is a number of echoes with substantial enough energies to require their suppression. Generally, for signal levels well above Johnson noise, most of the noise growth in the feed-forward FIR filtering process is attributable to the weighted terms not being derived from an echo-free signal, but rather from the signal supplied for equalization. The signal supplied for equalization usually includes echoes, of course. Accordingly, the weighted terms derived from that signal include not only a weighted principal component signal, but also a weighted echo spectrum. The weighted echo spectra combine in the feed-forward FIR filtering response to provide a principal source of noise growth.

In the design for a time-domain equalizer, an IIR filter cannot cancel pre-echoes. In some designs the IIR filter is also incapable of canceling short-delay post-echoes, because of latent delay in the filtering hardware. An IIR filter does not generate repeat echoes when it cancels (actually attenuates) a post-echo, and so it needs fewer weighting coefficients to suppress each echo. This reduces the number of opportunities for noise growth as compared to using a feed-forward FIR filter to suppress the same post-echo(es). The fact that the IIR filter does not generate repeat echoes makes it possible to employ programmable bulk delays for locating small clumps of non-zero weighting coefficients within the IIR filter structure where they will be most effective in canceling post-echoes.

The IIR filter combines differentially delayed, weighted responses to its output signal to be fed back for combination with its input signal to generate that output signal as IIR filter response with suppressed post-echoes. Since the IIR filter response is at least partially de-echoed, the combining of differentially delayed, weighted responses to that response is less susceptible to noise growth than combining a similar number of similarly weighted terms dependent on the signal supplied for equalization. This is a further reason that noise growth is less likely when canceling post-echoes by IIR filtering, rather than by feed-forward FIR filtering.

Care must be taken to insure stability in an IIR filter. Instability usually manifests itself in the IIR filter having a self-generated tendency for ramping its output signal up or down, so that output signal does not respond appropriately to the input signal. One method for determining when an IIR filter is apt to become unstable is to sum the coefficients in the IIR filter.[2] If the coefficients do not sum to more than unity, stability is assured.

Equalizer performance during severe multipath reception is significantly affected by the manner in which the cross-over between FIR filtering and IIR filtering is made. See Section 6.13.6 "Passband Equalization."

In early DTV receivers the transversal filters used for time-domain equalization filtering had specified delay structures that were not altered during operation. Later on, it was proposed to use programmable bulk delays in the IIR filtering used for canceling post-ghosts, so that digital multipliers could be conserved. Newer designs for time-domain equalizers temporarily store equalizer input signal samples in a memory associated with a computer that performs the delaying, weighting, and combining functions associated with transversal filtering. The computer and its associated memory are constructed within a monolithic integrated circuit. Using a computer for equalization allows digital multiplication steps to be better allocated to suppressing the most significant multipath terms, so sparse-tap equalization methods are more easily put into practice. Using a computer for equalization allows the principal signal or "cursor" samples to be changed during the course of operation, which is of benefit when having to deal with fast-changing multipath reception conditions. Using a computer for equalization loosens real-time-processing restrictions, allowing the recycling of DFS signal samples for more efficient use of Wiener methods of adaptation, for example. Using a computer for equalization may permit some development in equalization methods without the need for re-design of receiver hardware.

5.5 Considerations Concerning the Use of Decision-Feedback Equalization and the use of Linear-Feedback Equalization

5.5.1 Definitions of Decision-Feedback Equalizer (DFE) and Linear-Feedback Equalizer (LFE)

When IIR transversal filtering used for equalization applies its response as input signal to decision-making apparatus, the decision made by the decision-making apparatus can be re-coded in accordance with the symbol coding used in the original transmission and used as feedback signal instead of the IIR transversal filtering response itself. An equalizer using this type of IIR transversal filtering is called a "decision-feedback equalizer" (DFE).[3]

An equalizer called a "linear-feedback equalizer" (LFE) includes IIR transversal filtering that uses the equalizer response itself as feedback signal. It is a species of "linear equalizer" (LE). Equalizers that rely solely on FIR filtering are another species of linear equalizer.

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[2] C. B. Dietrich: "IIR Ghost Canceling System with Reduction of Filter Instability", U. S. patent No. 5 335 020, 2 August 1994.

[3] R. J. Keeler: "Construction and Evaluation of a Decision Feedback Equalizer", *Rec. IEEE Int. Conference on Communications*, Philadelphia, PA, pp. 19–21, June 1972.

5.5.1.1 Reasons for Using Decision-Feedback Equalization

Constraining the fed-back response of the equalizer to valid symbols without accompanying noise suppresses tendency for the IIR filter to ramp its response up or down independently of input signal to the filter.

A channel with steep-slope Nyquist roll-off benefits substantially from DFE providing that the decisions are error-free and timely available.[15.1] There is a more gradual increase in error rate as the signaling rate approaches the Nyquist limit. The 6 MHz DTV channel has a steep-slope Nyquist roll-off.

Assuming there are no decision errors, the decision-feedback signal tends to decrease the overall noise level of the output signal from the equalizer filter, because less noise is fed back to add in random phase to the noise from the feed-forward FIR filter response. This tendency is most evident when the signal supplied for equalization is accompanied by moderate amounts of band-limited Johnson noise, or multipath distortion, or both band-limited Johnson noise and multipath distortion. If the amount of noise and intermodulation distortion is large enough to cause decision errors much of the time, this noise improvement is no longer available.

Decision feedback is particularly useful in connection with adaptive equalization using data-directed decision-making apparatus. There is reduction in the digitized noise fed back in the IIR transversal filtering when SNR of the received signal is somewhat above the value where data slicing consistently fails. This reduces BER and improves the speed of convergence of adaptive equalizer weighting coefficients based on least-squared-error methods.

The DFE, being recursive, tends to stretch decision errors in time. Decision errors are apt to occur when impulse noise occurs in the channel or when Johnson noise keeps the SNR of the channel consistently low. The feeding back of erroneous decisions is most likely to cause to give rise to running errors when the post-echoes to be suppressed are not substantially weaker in energy than the principal (cursor) signal. The individual post-echoes most likely to be strongest relative to the principal signal are short-delay post-echoes. Accordingly, in some equalizer designs short-delay post-echoes are suppressed by feed-forward FIR filtering, rather than by IIR filtering. This permits the effects of decision errors to be curtailed immediately after trellis decoding determines the existence of those errors with a sufficiently high degree of confidence. The feed-forward FIR filtering structure is such that past equalizer responses can be corrected immediately after decision errors are determined with the requisite degree of confidence, with the correction being done by feed-forward means.

Insofar as data recovery is concerned, the stretched decision errors caused by an isolated occurrence of burst noise are accommodated by data-deinterleaving and Reed-Solomon error correction. Insofar as the adaptation of equalizer coefficients is concerned, the errors in equalizer response measured during each occurrence of impulse noise and the ensuing stretch interval are preferably replaced by zero-valued reception errors.

5.5.2 Latency Problems in Decision-Feedback Equalization (DFE)

A common problem with using DFE with coded systems is that short-delay or delay-free decisions are required to cancel Inter-Symbol Interference (ISI) components generated by the most recent symbols. Frequently, short-delay post-echoes are the most consequential ISI. An ordinary Viterbi trellis decoder has a minimum latency of four to five times the constraint length of the trellis code, before the decisions that the decoder furnishes are consistently reliable.[4] If

4 R. D. Gitlin, J. F. Hayes, and S. B. Weinstein: “7.5.3 Performance Comparison between Linear and Decision Feedback Receivers”, *Data Communications Principles*, pg. 506, pg. 551, Plenum Press, NYC and London, 1992.

such a trellis decoder is used in a decision-feedback loop, the short-delay or delay-free decisions extracted in real time are preliminary in nature. It has been reported that the preliminary decisions are often highly unreliable and if used for DFE can seriously degrade the performance of the decoder.[5]

However, except just after an impulse noise occurrence, preliminary trellis-decoding decisions provide better DFE than decisions based on simple data slicing. A receiver that suppresses impulse noise occurrences can be designed to base DFE on simple data slicing during the symbol epochs immediately following an impulse noise occurrence.

Quick or preliminary trellis-decoding decisions can be more reliably generated by so-called “smart” data slicers.[6] These “smart” data slicers use the properties of trellis coding to predict certain of the bits that will be coded in the symbol next to be received. This facilitates selection of decision-feedback signal from the responses of binary data slicers for set-partition or coset codes within the trellis coding. The decision is available in real time.

While symbol decisions that are generated quickly are very desirable for decision-feedback purposes, particularly for suppressing shorter-delayed post-echoes, symbol decisions that are as reliable as possible are more desirable for adaptation of equalizer weighting coefficients. More reliable symbol decisions permit larger step size in gradient-following adaptation algorithms, which speeds up adaptation despite some increase in the time taken for individual symbol decisions to be made. Accordingly, DTV receiver designs are apt to use a different symbol decision procedure for adapting equalizer weighting coefficients than they use for generating decision-feedback signal for the recursive filtering that suppresses post-echoes.

The cascading of NTSC co-channel interference suppression filtering and associated ISI-suppression filtering after the equalization filtering creates latent delay in the decision-feedback loop, which latent delay is typically only two symbol epochs or so in duration. In some designs NTSC co-channel interference suppression filtering precedes the equalization filtering in cascade to avoid introducing delay into the decision-feedback loop. In other designs NTSC co-channel interference suppression filtering is done by feed-forward filtering methods to avoid introducing delay into the decision-feedback loop.

If the decision-feedback equalizer is a fractional equalizer, the decimation filtering before data slicing and the interpolation filtering used in re-sampling the decisions to the sampling rate used in the fractional equalizer can introduce latent delay into the decision-feedback loop. In a T/2 fractional equalizer this latent delay can be kept to a single symbol epoch duration.

If the equalizer is a passband equalizer, the VSB AM modulator used for converting decision feedback to passband frequencies will introduce fairly lengthy latent delay attributable in major part to the digital vestigial-sideband filtering. This problem is avoided by using an equalizer design comprising a passband feed-forward FIR filter and a baseband FIR filter for decision feedback.[7]

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[5] J. J. Nicolas and J. S. Lim: “Rapid-Update Adaptive Channel-Equalization Filtering for Digital Radio Receivers, such as HDTV Receivers”, U. S. patent No. 5 453 797, 26 September 1995, col. 9, lines 19–28.

[6] S. N. Hulyalkar, T. J. Endres, T. A. Schaffer, and C. H. Strolle: “Method of Estimating Trellis Encoded Symbols Utilizing Simplified Trellis Decoding”, U. S. patent No. 6 178 209, 19 June 1998.

[7] C. H. Strolle and S. T. Jaffe: “Blind Equalizer Method and Apparatus for HDTV Transmission Using an NTSC Rejection Filter for Mitigating Co-Channel Interference”, U. S. patent No. 5 550 596, 24 November 1998.

5.5.3 Parallel Decision-Feedback Equalization

The latency problem with decision-feedback equalization of baseband DTV signal can be overcome using parallel decision-feedback equalization.[8] The response of a feed-forward FIR filter is supplied as input signal to a plurality of IIR filters employing decision feedback. There is a respective IIR filter for each of the survivor paths in the Viterbi trellis decoder.

The adaptation of the weighting coefficients of the feed-forward FIR filter and the plurality of IIR filters employing decision feedback are done in the following way. The final hard-decision of the trellis decoder response as to the value of each successive symbol is compared with the soft-decision response the decision-feedback IIR filter generating the final hard-decision supplies, to generate an equalization error signal. This equalization error signal is used by an auto-regression algorithm for generating adjustments of the weighting coefficients of the feed-forward FIR filter and of the decision-feedback IIR filter generating the final decision. The weighting coefficients of each of the other decision-feedback IIR filters are adjusted to correspond to the weighting coefficients of the decision-feedback IIR filter generating the final decision. So are the weighting coefficients of the linear-feedback IIR filter.

5.5.4 Reasons for Using Linear-Feedback Equalization (LFE)

Usually the feedback loop in an LFE can be closed with no more than a sample epoch delay. Short-delay post-echoes are better suppressed by LFE IIR filtering than by the feed-forward FIR filter. Suppressing these short-delay post echoes by the feed-forward FIR filter generates repeat post-echoes with extended delay, which undesirably tends to lengthen the overlap of FIR and IIR filtering in the time domain that is needed for good post-echo suppression. Suppressing short-delay post echoes by LFE generates no repeat echoes, and it can be used together with decision-feedback suppression of longer-delay post-echoes.

A principal reason for using linear feedback in the IIR filter portion of an equalizer is that it permits the IIR filter to precede the feed-forward FIR filter in cascade. The IIR filter suppresses post echoes without generating repeats. Since the feed-forward FIR filter suppresses echoes using differentially delayed and scaled responses to its input signal and its accompanying echoes, the previous suppression of post-echoes in that input signal reduces the incidence of repeats in the post-echo direction in the feed-forward FIR filter response. The feed-forward FIR filter can be used to suppress pre-echoes without generating post-echoes of consequence, so there is less (if any) need of overlap of FIR and IIR filtering in the time domain for good post-echo suppression. This can save some need for multiplication capability in the equalizer.

Least-mean-squares adaptation of a linear equalizer minimizes the joint effects of Gaussian noise and ISI. Some ISI is tolerated in order to reduce the noise enhancement.

In contrast to the DFE, the LFE can employ fractional equalization in its IIR filtering as well as in its feed-forward FIR filter. Fractional equalization in the IIR filtering permits the suppression of post-cursor ISI, no matter whether it is in phase with symbol epochs in the principal signal or is slipped in phase with respect to those symbol epochs. This can aid symbol synchronization.

LFE can be used during initialization of the equalizer weighting coefficients with switch-over to DFE after they have been determined. This avoids initialization being hampered by the running errors associated with DFE using decisions based on trellis decoding.[9]

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[8] R. D. Gitlin and N. A. Zervos: "Decision Feedback Equalization with Trellis Coding", U. S. patent No. 5 056 117, 8 October 1991.

[9] S. N. Hulyalkar, T. J. Endres, R. A. Casas, and C. H. Strolle: "On Equalizer Design for Digital Television Channels in the United States", 2000 IEEE Broadcasting Symposium, Washington, D.C., September 2000.

5.5.5 Combined Linear-Feedback Equalization and Decision-Feedback Equalization

An LFE that cascades the feed-forward FIR filter after the linear-feedback IIR filter supplies its response as a first input signal to a linear combiner, the response of which linear combiner is applied as input signal to subsequent decision-feedback IIR filtering. The input signal applied to the linear-feedback IIR filter is differentially combined with the response of the linear-feedback IIR filter to separate the post-echo ensemble accompanied by noise generated during the linear-feedback IIR filtering. The post echo ensemble is delayed to compensate for the latent delay in the feedfoward FIR filter. The delayed post echo ensemble is supplied as a second input signal to the linear combiner, the combined first and second input signals to the linear combiner in effect being the equivalent of the input signal to the LFE with pre-echoes suppressed by the feed-forward FIR filtering. This feed-forward FIR filtering has the benefit of suppressing the pre-echoes with its input signal being free of post-echoes. Accordingly, there is no need for extending the time-domain response of the feed-forward FIR filter into the post-echo region.

The decision-feedback IIR filtering, which can comprise parallel DFEs for each survivor in the Viterbi trellis decoding, cancels the post-echoes from its response to the output signal the linear combiner supplies. The final hard-decision of the trellis decoder response as to the value of each successive symbol is compared with the soft-decision response that the decision-feedback IIR filter generating the final hard-decision supplies. This comparison generates an equalization error signal. This equalization error signal is used by an auto-regression algorithm for generating adjustments of the weighting coefficients of the feed-forward FIR filter and of the decision-feedback IIR filter generating the final decision. The weighting coefficients of each of the other decision-feedback IIR filters are adjusted to correspond to the weighting coefficients of the decision-feedback IIR filter generating the final decision. So are the weighting coefficients of the linear-feedback IIR filter.

This type of combined equalizer has the benefit of the faster convergence of weighting coefficients that decision-feedback provides when a Kalman adaptation technique is employed. However, the generation of repeat post-echoes in the FIR feed-forward FIR filter is avoided, helping to control noise growth that occurs when the principal reception component has little more energy than the echo spectrum. However, the improvements over types of DFE that do not use a linear-feedback IIR pre-filter are significant only for exceptionally difficult multipath reception conditions encountered at very small fraction of reception sites. While the number of multiplications required in the feed-forward FIR filter is reduced because the filter kernel need not overlap the post-echo region appreciably, the number of multiplications in the IIR filtering is increased, owing to the use of the linear-feedback IIR filtering in addition to the decision-feedback IIR filtering.

5.6 Issues Concerning the Nature of Sampling in the Equalizer

5.6.1 Nyquist-Rate Clocking of the Equalizer

The Grand Alliance prototype DTV receiver [3] uses equalization filtering that is clocked at the 10.76 megasamples/s Nyquist rate. The equalizer is a synchronous equalizer with a kernel taps for weighting coefficients at symbol-epoch intervals.

Nyquist-rate clocking of the equalizer does not provide the “excess” bandwidth required to perform symbol synchronization subsequent to equalization by stochastic methods. However, phase discrimination of the DSS symbols in the equalizer response which change at half Nyquist rate can develop an error signal for adjusting the phase of the Nyquist-rate sampling of the equalizer input signal to best suit the Nyquist-rate clocking of the equalizer and the data slicer that follows. In the Grand Alliance prototype DTV receiver such an error signal is used for adjusting the phase of the IF DTV signal that is synchronously detected.

5.6.2 The Avoidance of Under-Sampling During Equalization

Ideally, the data slicer should receive an input signal with a system response that has a frequency spectrum that exhibits a Nyquist slope—a raised-cosine roll-off of energy 6 dB down at 5.38 MHz, half the 10.76 MHz Nyquist frequency. If the tail of the raised-cosine roll-off above 5.38 MHz is not preserved to frequencies where its energy is inconsequential, there will be increased ISI. Nyquist-rate clocking of the equalization filtering at 10.76 megasamples/s undersamples the tail of the raised-cosine roll-off above 5.38 MHz, causing aliasing that increases ISI and raises the symbol error rate during data slicing of baseband DTV signals with SNRs close to TOV.

For best performance, the equalizer should be clocked so as to provide adequate bandwidth to sample the tail of the raised-cosine roll-off that can extend up to 5.7 MHz. Symbol synchronization and re-sampling to symbol rate require equalizer response with preserved raised-cosine roll-off for the symbols in the re-sampled signal to be independent of each other. Independence of the symbols in the re-sampled signal improves the performance of stochastic gradient techniques for adapting equalizer parameters.

If the clocking rate through the equalizer is twice symbol rate, re-sampling to symbol rate can be done simply with a symbol synchronizer using linear interpolation. Higher-order interpolation does not appear to provide significant advantage over linear interpolation.

5.6.3 Synchronous Equalization vs. Fractional Equalization

In synchronous equalization filtering, the kernel taps provided with respective weighting coefficients are spaced a symbol epoch τ apart. The clocking rate through the filter is some multiple of Nyquist rate. Clocking the synchronous equalization filtering at twice Nyquist rate facilitates symbol synchronization and avoids ISI that would be caused by undersampling baseband DTV frequencies in the 5.38–5.69 MHz region.

In fractional equalization filtering, the kernel taps provided with respective weighting coefficients are spaced a fraction of a symbol epoch τ apart. This requires that the clocking rate through the filter be some multiple of Nyquist rate, multiplied by the reciprocal of the fraction. The clocking rate can be chosen to be the Nyquist rate multiplied by the reciprocal of the fraction. The usual practice in fractional equalization is to space the weighting coefficients half a symbol epoch apart with the equalization filtering being clocked at twice Nyquist rate.

Fractional equalization clocked at a multiple m times Nyquist rate and provided with weighting coefficients at kernel taps spaced τ/m apart can suppress an echo more precisely in time without involving as many of the weighting coefficients. The fewer weighting coefficients involved in suppressing an echo, the better the ISI from echoes out of phasing with optimal symbol phasing is avoided. The cost of this better ISI suppression is that the number of multiplication procedures involved in equalization increases by the factor m. The factor m is seldom chosen larger than two in DTV receivers for the consumer market.

Fractional equalization clocked at a twice Nyquist rate and provided with weighting coefficients at kernel taps spaced $\tau/2$ apart suppresses essentially completely the ISI from echoes in quadrature phasing with optimal symbol phasing. This removes a significant source of "noise" from the data slicing process during certain multipath distortion conditions.

An equalizer clocked at higher than Nyquist rate has to be followed by a decimation filter that decimates the equalizer response to symbol rate. This re-sampling to symbol rate is a pre-requisite of data slicing, and the decimation filtering can be incorporated within symbol synchronizing that precedes the data slicing. The equalizer clocked at higher than Nyquist rate can employ IIR filtering with decision feedback. Symbol coding of data slicing decisions generates samples of decision feedback signal at symbol rate. The ways in which the decision feedback signal is utilized are different in a synchronous equalizer clocked at Nyquist rate, in a synchronous equalizer clocked at a multiple of Nyquist rate, and in a fractional equalizer.

If the feed-forward FIR filter provides synchronous equalization and is clocked at Nyquist rate, the IIR filtering with decision feedback is connected in cascade after the feed-forward FIR filter. The equalization error signal for use in Kalman adaptation procedures is generated by comparing the decision feedback signal with the response of the IIR filtering with decision feedback, as delayed to compensate for the latent delay in the symbol decoding and re-coding procedures used to generate the decision feedback signal.

If the feed-forward FIR filter provides synchronous equalization, but is clocked at higher than Nyquist rate and is followed by a decimation filter, it is simplest to cascade the IIR filtering with decision feedback after the decimation filter and to clock that IIR filtering at Nyquist rate. More particularly, the IIR filtering with decision feedback comprises a feedback FIR filter clocked at Nyquist rate for processing decision feedback samples and a linear combiner for combining the Nyquist-rate feedback FIR filter response with the Nyquist-rate decimation filter response to the over-sampled feed-forward FIR filter response. The equalization error signal for use in Kalman adaptation procedures is generated by comparing the decision feedback signal with the response of the IIR filtering with decision feedback, as delayed to compensate for the latent delay in the symbol decoding and re-coding procedures used to generate the decision feedback signal.

The preferred procedure for minimizing ISI in a fractional equalizer is to use the same over-sampling clock rate in the IIR filtering with decision feedback as in the feed-forward FIR filtering.[10] Symbol decisions are symbol coded and re-sampled to that oversampling clock rate to provide a decision-feedback signal applied as input signal to the feedback FIR filter. The responses of the feed-forward FIR filter and the feedback FIR filter are combined to generate a fractional equalizer response sampled at the over-sampling clock rate. This response is applied as input signal to the decimation filter, which supplies decimated fractional equalizer response at Nyquist rate to the data slicer. The equalization error signal used for adaptation of the weighting coefficients of the fractional equalizer is generated by comparing the fractional equalizer response, as delayed to compensate for the latent delay in the generation of decision-feedback signal from that response, to the decision-feedback signal. Both signals involved in this comparison are sampled at the over-sampling clock rate, so the equalization error signal has the temporal resolution required for generating independent weighting coefficients for τ/n kernel tap spacing in the feed-forward FIR filter and in the feedback FIR filter.

Equalization filtering that cascades synchronous IIR filtering with decision feedback after fractional feed-forward FIR filtering is described in textbooks. The equalization error signal used for adaptation of the weighting coefficients of this type of equalization is generated by comparing the synchronous IIR filtering response, as delayed to compensate for the latent delay in the generation of decision-feedback signal from that response, to the decision-feedback signal. This equalization error signal has the temporal resolution for generating independent weighting coefficients for synchronous equalization, but not for fractional equalization. Intermediate weighting coefficients for the fractional feed-forward filtering are generated by interpolation. This constrains this form of adaptive equalization filtering to provide no better equalization than

[10] A. L. R. Limberg and C. B. Patel: “Adaptive Fractionally Spaced Equalizer for Received Radio Transmissions with Digital Content, Such As DTV Signals”, U. S. patent No. 6 377 312, 23 April 2002.

can be achieved using an adaptive synchronous equalizer clocked at the same oversampling rate and provided the same kernel span. Such a synchronous equalizer requires only half as many multiplications in the feed-forward FIR filter. So, if IIR filtering with decision feedback is used in a fractional equalizer, the feed-forward FIR filter and the feedback FIR filter should be clocked at the same over-sampling rate with decision-feedback signal being re-sampled to that over-sampling rate for application to the feedback FIR filter as input signal.

5.6.4 Real-Only Equalization vs. Complex Equalization

Single-sideband amplitude-modulation (SSB AM) signals can be generated by combining double-sideband amplitude-modulated (DSB) in-phase (I) carrier and DSB AM quadrature-phase (Q) carrier offset 90 degrees from the I carrier, the modulating signals of which DSB AM signals are related by Hilbert shift. In a Hilbert shift all frequency components of a signal are shifted 90º in phase, and a filter performing such phase shifting is referred to as a Hilbert filter. Customarily, a data stream modulates the amplitude of the I carrier and Hilbert filter response to the data stream modulates the amplitude of the Q carrier. In the SSB receiver, complex demodulation recovers the data stream with accompanying real echoes of the reproduced baseband signal and the Hilbert filter response to the data stream with accompanying imaginary echoes of the reproduced baseband signal. If equalization is done in the IF passband, a complex equalizer has to be used. If the SSB receiver uses baseband equalization, both of the two orthogonal components of the complex demodulation result can be equalized in a procedure known as *complex equalization*. If the baseband equalization is included in the loop used for carrier recovery, a complex equalizer has to be used.

Otherwise, however, the baseband equalization can be done using only one of the two orthogonal components of the complex demodulation result, since the two components do not contain independent information. If only one of the two orthogonal components of the complex demodulation result is equalized, it is customary to equalize the real-only component in which the data stream appears directly. If the SSB signal is accompanied by a pilot carrier in phase with the suppressed I carrier, the imaginary-only component of the complex demodulation result in which the pilot carrier is suppressed can be Hilbert filtered and equalized instead. Indeed, adaptive equalization can automatically configure the equalization filtering to perform the Hilbert filtering as well as suppressing echoes, without need for a preceding Hilbert filter.

The 8-VSB (also 16-VSB) system uses vestigial-sideband amplitude modulation, rather than single-sideband amplitude modulation, but the two types of modulation are similar except close to carrier frequency. Baseband equalization of 8/16 VSB signals can be performed on just one of the two orthogonal components of the complex demodulation result.

A complex equalizer with weighted kernel taps at symbol-epoch intervals can provide similar performance to a fractional equalizer with weighted kernel taps at half-symbol-epoch intervals. There have been attempts to use this fact to reduce clocking rate through the equalizer to symbol rate rather than twice symbol rate, but clocking at symbol rate undesirably undersamples the tail of the raised-cosine roll-off of the baseband DTV signal above 5.38 MHz.

If the complex equalizer operates on I-channel and Q-channel demodulation results, it requires four times the number of weighted kernel taps that are in a synchronous real-only equalizer with equivalent echo-suppression range. Instead of just an adaptive filter for the I-channel signal, there are also an adaptive filter for the Q-channel signal, an adaptive filter for the

I-channel signal crosstalk into the Q-channel signal, and an adaptive filter for the Q-channel signal crosstalk into the I-channel signal.[11]

From the standpoint of the number of weighted kernel taps required at symbol-epoch intervals, a better way to provide equalized I-channel and Q-channel signals is to demodulate at +45° and +135° axis and use a respective adaptive filter for each of the demodulation results. The adaptive filter responses are then differentially combined to generate the equalized I-channel signal and additively combined to generate the equalized Q-channel signal. That is, twice the number of weighted kernel taps at symbol-epoch intervals are required as in a synchronous real-only equalizer with equivalent echo-suppression range. This is the same number of weighted kernel taps as in a fractional equalizer with taps at half-symbol-epoch intervals and with equivalent echo-suppression range.

5.6.5 Equalization Within the Carrier Recovery Loop vs. Equalization Independent of the Carrier Recovery Loop

The reason for considering inclusion of equalization in the carrier recovery loop used for AFPC of local oscillations used for synchrodyning is that demodulation is optimized for the principal component of received signal. Carrier recovery is then, at least in theory, less affected by changes in the echo components of the received signal.

If the equalizer is included in the carrier recovery loop, it must be a complex equalizer. This is the case even if equalization is done at baseband. Equalization has then to be provided for the imaginary component of the baseband DTV signal, which is used in developing automatic frequency and phase signal to control the local oscillations used for synchrodyning the DTV signal. This is besides the equalization provided for the real component of the baseband DTV signal, which is used in reproducing data and is additionally used in a carrier recovery loop that employs a Costas technique.[12]

If the equalizer is included in the carrier recovery loop, the zero-frequency pilot component of the complex baseband DTV signal must be preserved during equalization. Including the equalizer in the carrier recovery loop increases open-loop delay, typically by 60 to 100 microseconds. This makes it difficult to design a carrier recovery loop that slews fast enough to track phase noise, but presents no risk of being self-oscillatory. An equalizer that occasionally shifts cursor position (i.e., the latent delay for the principal component of the received signal) is poorly suited for inclusion in the carrier recovery loop.

Based on system simulations, it appears that the better system design is not to include equalization in the carrier recovery loop, nor to try to track phase noise during demodulation. Instead, adaptive complex equalization is performed outside the carrier loop to suppress echoes that are appreciably advanced or delayed respective to the principal signal. Then, the constellation of complex equalization results is de-rotated to degenerate the phase modulation introduced by phase noise in the local oscillators and by echoes that are little advanced or delayed respective to the principal signal. This de-rotation can be done using stochastic techniques, checking to see what degree of de-rotation results in the least amount of decision error.

[11] D. S. Han: "Apparatus and Method for Controlling Operation of a High Defination [sic.] Television Adaptive Equalizer", U. S. patent No. 5 661 528, 26 August 1997.

[12] J. P. Costas: "Synchronous Detector System", U. S. patent No. 3 101 448, 23 December 1964.

5.7 Treatment of Pilot in Equalization Filtering

5.7.1 Suppression of Pilot When Computing DFT of Received DTV Signal

The usual practice is to suppress the pilot accompanying a received DTV signal before computing the DFT of the received signal. The pilot interferes with the constancy of spectral energy down to zero-frequency in the recovered baseband signal. This is the reason for its suppression when performing equalization in the frequency domain using techniques that equalize energy in the DFT terms.

In a procedure for initially determining the weighting coefficients of a time-domain adaptive equalizer, the DFT of a training signal portion of the received signal is computed for term-by-corresponding-term division by the DFT of an ideal training signal to generate the DFT characterizing the channel. The inverse-DFT of the DFT characterizing the channel is the channel impulse response (CIR), from which the initial weighting coefficients are computed. This procedure for obtaining the CIR corresponds to deconvolution in the time domain. The DFT of the ideal training signal is stored in read-only memory within the receiver. The DFT is more simply stored in ROM if the DFT does not have to take into account a large zero-frequency component attributable to demodulated pilot.

5.7.2 Treatment of Zero-Frequency Pilot in Baseband Equalization Filtering

The synchrodyning of IF DTV signal to baseband generates a zero-frequency or direct term in the baseband DTV signal as demodulated, which term is attributable to demodulation of the pilot. It is desirable to suppress the zero-frequency pilot in the equalized baseband DTV signal supplied for data slicing in the trellis decoder or other quantizing apparatus. Suppressing the zero-frequency pilot removes a source of uncertainty as to the boundaries between data slicing bins.

If a baseband equalization filter response is used in the carrier recovery loop, the zero-frequency pilot must be preserved in that response. Accordingly, the zero-frequency pilot cannot be suppressed in the baseband DTV signal supplied as input signal to the baseband equalization filter. Instead, the zero-frequency pilot can be suppressed in the baseband equalization filter response supplied for data slicing in the trellis decoder or other quantizing apparatus, although it must be preserved in the baseband equalization filter response used in the carrier recovery loop.

If the equalized baseband DTV signal is supplied as the response of a baseband equalization filter that is separate from the carrier recovery loop, zero-frequency pilot can be suppressed in the baseband DTV signal supplied to that filter as input signal. This disposes the DTV signal more symmetrically around the zero-axis and therefore permits better use of the dynamic range in the baseband equalization filter, which is symmetrical around the zero-axis.

5.7.3 Pilot Suppression Circuitry

See Section 6.14 ("Treatment of Zero-Frequency Pilot Accompanying Baseband DTV Signal"), *infra*.

5.8 Multiple Echoes

The number of echoes to be canceled has in itself little effect on the theoretical performance of the system. Theoretical limits to successful echo suppression depend on the amount of noise growth that occurs in compensating the frequency response of the received signal and suppressing accompanying echoes. The quantization that is done in decision feedback largely avoids noise growth in recursive filtering portions of the receiver equalization filtering. (Refer to the description of decision feedback in Section 5.5.1.1.) Noise growth occurs primarily in the feed-forward FIR filter portions of the receiver equalization filtering, which delay and scale the received signal in various ways to generate signals that are combined with that received signal for suppressing the accompanying echoes. The echoes accompanying the principal component of the received signal as delayed and scaled in various ways do not cancel in these combining procedures but persist as repeat echoes that are components of noise growth. Pre-echoes and post-echoes that are nearly as large as the principal component of the received signal and that are suppressed by the feed-forward FIR filter portions of the receiver equalization filtering cause substantial noise growth. A large number of pre-echoes and post-echoes that are suppressed by the feed-forward FIR filter portions of the receiver equalization filtering cause substantial noise growth, even though individually they are not nearly as large as the principal component of the received signal. When noise growth reduces the SNR of the equalizer response below the threshold of visibility of video errors, the reception system as a whole fails to perform satisfactorily.

5.9 The Need for Adaptive Equalization Filtering in a Receiver for Over-the-Air DTV Broadcast Signals

The changing multipath reception conditions encountered in over-the-air DTV broadcasting using transmission antennas mounted on ground-based towers mandate that the equalizers in DTV receivers for receiving such broadcasting be adaptive in nature. The fairly high probability at many reception sites of appreciable change in multipath conditions every few milliseconds (so-called "dynamic" multipath) makes it desirable that the equalizer parameters be updated substantially more frequently than once every data field. Initialization of equalizer filtering parameters within a fraction of a second is desirable in a DTV receiver design for the consumer market, so the user of the receiver is less likely to find the response of the receiver to change in reception channel being too slow.

5.9.1 Choice of Kalman or Wiener Method(s) for Adaptive Equalization

A first approach to adaptive equalization uses data-driven adaptive filtering methods of the type originally described by R. E. Kalman, which methods will be generically referred to as "Kalman methods" in this document.[13] [14] Most designs of DTV receivers for the consumer market use a Kalman method for adapting equalizer parameters based on characteristics of the continuing flow of data. These methods permit the equalizer parameters to be updated frequently or even continuously. Kalman methods adapt the equalizer parameters by auto-regression that depends on the differences between symbols as received and estimates of the corresponding symbols that were transmitted. The estimates of the transmitted symbols are based on the symbols as received, without relying on specific foreknowledge of what symbols were actually transmitted. The estimates are based on foreknowledge of the general characteristics received symbols should have for a perfect reception channel. Estimates of the transmitted symbols are commonly based on data-slicing results, the results of partial trellis decoding, the results of complete trellis decoding, or some combination of such results.

The updating of equalizer filtering parameters by a Kalman method must be done incrementally, with the sizes of the incremental changes to the filtering parameters being kept small enough to avoid erroneous estimates of the transmitted signal having too much effect on the adaptation of the equalizer parameters. Accordingly, the initialization of equalizer filtering

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[13] R. E. Kalman: "A New Approach to Linear Filtering and Prediction Problems", *Trans. ASME J. Basic Eng.*", vol. 82, pp. 35–45, 1960.

[14] R. E. Kalman and R. S. Bucy: "New Results in Linear Filtering and Prediction Theory," *Trans. ASME J. Basic Eng.*", vol. 83, pp. 95–108, 1961.

parameters by a Kalman method can be slow, particularly until some tracking of the adaptation to changing multipath conditions can be established. Indeed, initialization may never be attained in a short enough time to be practical. So, initialization of equalizer filtering parameters using some auxiliary method is customary in most designs of DTV receivers for the consumer market.

Since the adaptation of equalizer filtering parameters by a Kalman method tends to be slow anyway, it is preferable to re-coded final trellis decoding results for estimating the errors in the equalizer response that are used for adaptation. Equalization errors are more accurately estimated from re-coded final trellis decoding results than from re-coded preliminary trellis decoding results or from data-slicing the equalizer response. This allows the incremental changes made to the filtering parameters each symbol epoch to be larger in size, speeding up convergence to compensate at least in part for the lag in obtaining final trellis decoding results.

A second approach to adaptive equalization uses methods of filter adaptation of the type originally described by N. Weiner, which methods rely on prescribed training signals, and will be generically referred to as "Wiener methods" in this document. Wiener methods adapt equalizer parameters by auto-regression that depends on the differences between symbols as received and the corresponding symbols as transmitted, based on the receiver having reliable foreknowledge of what symbols were actually transmitted. The error measurements based on these differences are more reliable than the differences between symbols as received and estimates made by a Kalman method of the corresponding symbols that were transmitted. So, the sizes of the incremental changes to the equalizer parameters can be made larger without having too much effect on the adaptation of the equalizer parameters. Consequently, if the training signal is sufficiently long, convergence of the equalizer parameters can be considerably faster than is possible using a Kalman method. The training signal length needed for a Wiener method to initialize equalization of very difficult echo ensembles has been measured to be about 3000 known symbols, providing no errors intervene.[15] This requirement is appreciably greater than the 704 known symbols at the beginning of a data field as specified in A/53 [2]. So, unless the 704 known symbols at the beginning of a data field are stored and continuously re-cycled during the data field through a twin of the adaptive equalizer, adaptation by Wiener methods is also slow.

Some DTV receiver designs use the first 704 symbols of each data field for adaptive equalization using a Wiener method, and the first four symbols of the second through 313th data segments of each data may also be employed in adaptive equalization by a Wiener method. In early DTV receiver designs Wiener methods were used for initializing adaptive equalizer parameters, but "snapshot" initialization using just the first 704 symbols of a single data field is usually impossible unless there is quite accurate foreknowledge of the required equalizer parameters. So, relying solely on an intermittently performed Wiener method, convergence of equalization parameters to proper values can take several frame times for multipath reception having only static echoes. Convergence for multipath reception having dynamic echoes with substantial energy was seldom possible relying on just a Wiener method alone. With better initialization methods now being known, a common practice is to use a Wiener method for improving tracking of data-based adaptive equalization, but not for initializing adaptive equalizer parameters. This is described in Section 5.12.3, *infra*.

5.10 Choice of Algorithm for Adapting Equalizer Parameters

There are several auto-regression algorithms available that can be used for adapting the filter coefficients. A relatively simple algorithm that is widely used is the least mean squares (LMS)

[15] C. B. Patel, A. L. R. Limberg, and J. D. McDonald: PLM Presentation Before ATSC T3/S9, Washington D.C., June 7, 2001 (slide 9).

method.[16] The adaptive equalizer in the Grand Alliance prototype receiver used the LMS algorithm.

When the equalizer is first started, usually the initial tap weights do not compensate for the channel distortions adequately. In order to force initial convergence of the equalizer coefficients, a training signal known to both the transmitter and receiver is used as the reference signal. The error signal is generated by subtracting a locally generated copy of the training signal from the response of the adaptive equalizer. When using the training signal, the eye diagram is typically closed. The training signal serves to open the eye. After adaptation with the training signal, the eye has opened, and the equalizer may be switched to a decision-directed mode of operation. The decision-directed mode uses the symbol values at the output of the decision device instead of the training signal to estimate the transmitted DTV signal.[10] The estimation of the transmitted DTV signal is subtracted from the response of the adaptive equalizer to generate the error signal used for adapting the equalizer parameters. Decision-directed techniques are particularly useful when the channel impairments change during the data field.

A problem arises in the above scenario when a training signal is not available. In this case, a method called *blind equalization* is used to acquire initial convergence of the equalizer taps and force the eye open. In blind equalization decision-directed methods of adaptive equalization are used alone, without relying on a training signal or previous procedures for initializing equalizer parameters to substantially correct values. Blind equalization is unable to establish the initial weighting coefficients of an adaptive equalizer unless multipath reception conditions are quite benign, so decision-directed methods of adaptive equalization are better used in combination with other methods to establish the initial weighting coefficients of the adaptive equalizer.

Blind equalization has been extensively studied for quadrature amplitude modulated (QAM) systems. For QAM systems, there are several methods typically employed. The constant modulus algorithm (CMA) and the reduced constellation algorithm (RCA) are among the most popular.[17] [18] Modifications of these algorithms suitable for VSB systems are known and are described in Section 6.13.1.6 ("Initialization of Adaptive Equalizer Coefficients using Constant Modulus Algorithm (CMA)") and Section 6.13.1.7 ("Initialization of Adaptive Equalizer Coefficients using Modified Reduced Constellation Algorithm (MRCA)"), *infra*.

The recursive least squares (RLS) algorithm is an alternative decision-directed method for adaptive equalization that converges about 10 times faster than the LMS algorithm.[17] The convergence rate of the LMS algorithm is dependent on the channel characteristics and is even slower on channels which contain spectral nulls. The convergence rate of the RLS algorithm is not affected by the channel characteristics, so it is consistently fast.[19] Unfortunately, the number of computations used in the RLS algorithm is squared with increase in the number of equalizer data taps, rather than increasing linearly with that number as in the LMS algorithm.

Variants of the RLS algorithm are more modest in their computational requirements. The FTF algorithm employs four separate transversal filters responsive to the same data to generate the solution to the recursive-least-squares (RLS) problem. The first filter performs recursive

[16] Shahid U. H. Qureshi: "Adaptive Equalization," *Proceedings of the IEEE*, vol. 73, no. 9, pp. 1349–1387, September 1985.

[17] Walter Ciciora, et al.: "A Tutorial on Ghost Canceling in Television Systems," *IEEE Transactions on Consumer Electronics*, vol. CE-25, no. 1, pg. 9–44, February 1979.

[18] Gottfried Ungerboeck: "Fractional Tap-Spacing Equalizer and Consequences for Clock recovery in Data Modems," *IEEE Transactions on Communications*, vol. COM-24, no. 8, pp. 856–864, August 1976.

[19] Chun-Ming Zhao, Xiao-Yu Hu, and Xiao-Hu Yu: "Block Sequential Least Squares Decision Feedback Equalization Algorithm with Application to Terrestrial HDTV Transmission," *IEEE Trans. on Broadcasting*, vol. 44, no.3, pp. 278–285, September 1998.

forward prediction. The second filter performs recursive forward prediction. The third filter performs recursive computation of the gain vector. And the fourth filter performs recursive estimation of the desired response. The combination of the responses of these four filters produces the exact solution of the RLS problem, with a computational cost that increases linearly with the number of taps in the equalizer.[20] About 3 to 4 times as many computations are required as for the LMS algorithm, but word truncations are not well tolerated over extended time. Consequently, periodic correction of equalizer coefficients using another method is advisable.

Another species of the RLS algorithm, referred to in the literature as the Fast Kalman algorithm, converges appreciably faster than the LMS algorithm, but has only 7 to 8 times as many computations.[13] However, the Fast Kalman algorithm is problematic because it exhibits high sensitivity to round-off errors and other minor errors.

5.10.1 The Data Field Sync (DFS) Signal

The VSB signal contains features that facilitate the design of receivers that reliably perform the functions of acquiring and locking to the transmitted signal. The Data Field Sync (DFS) signal depicted in Figure 5.15 is a unique type of Data Segment in the VSB signal, and is used as the initial segment of each data field. Equalizer training signals composed of pseudo-noise (PN) sequences are major parts of the DFS signal. All payload data in the VSB signal is contained in the second through 313th data segments of its data fields, which data segments except for the Data Segment Sync (DSS) sequences are processed with data interleaving, Reed-Solomon error coding, and trellis coding. The segment containing the DFS signal is not processed this way, however, so the DFS signal is suited for measurement of linear distortion in the channel so that distortion can be compensated for. Furthermore, the Data Segment Sync (DSS) portions of the Data Segments are not processed with data interleaving, Reed-Solomon error coding, and trellis coding.

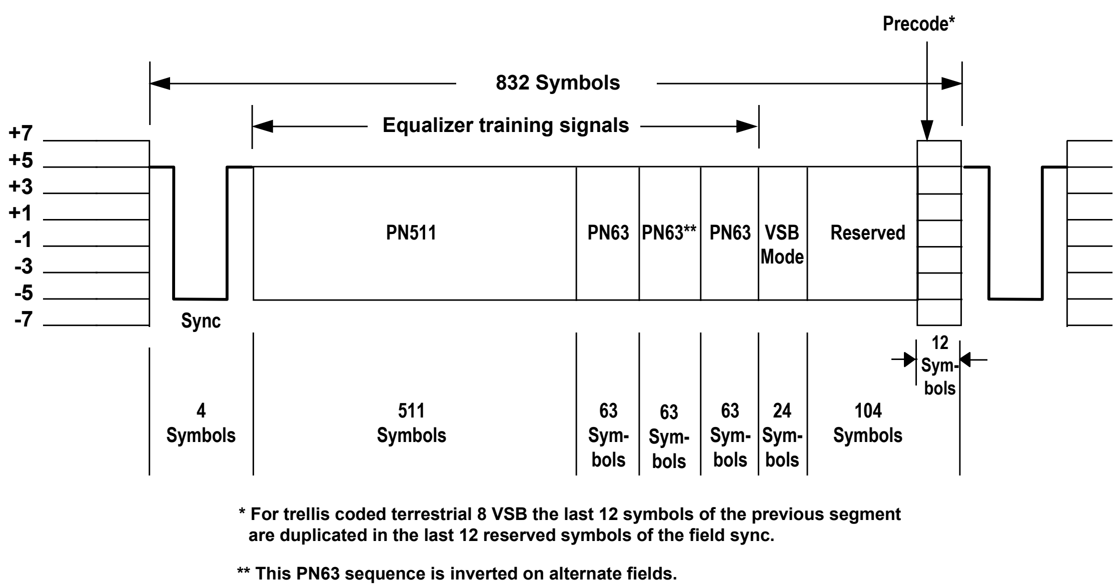

Figure 5.1 Data field sync.

The equalization of the signal for suppressing echoes and otherwise correcting channel frequency response is facilitated by the inclusion of prescribed symbols in the Data Field Sync

[20] S. Haykin: "Fast Transversal Filters", Chapter 16, *Adaptive Filter Theory*, Prentice Hall, 1991.

signal. These prescribed symbols comprise the four symbols in the DSS sequence beginning the initial data segment of the data field, a PN511 sequence consisting of 511 symbols immediately following that initial DSS sequence, and a triple PN63 sequence consisting of 189 symbols immediately following the PN511 sequence.

The Grand Alliance receiver [3] used the Segment Sync first to acquire and then to synchronize to the VSB signal. The Grand Alliance receiver subsequently used the known symbols in the DFS signal both to identify the reception of those known symbols and then to utilize those known symbols for equalization of the channel to reduce linear transmission distortions. One of the ways that adaptive equalization can be done in the Grand Alliance prototype receiver design employs a Wiener method for adjusting the weighting coefficients responsive to the measured distortion of the known symbols in the DFS signal.

In alternative receiver designs that use bright-spectral-line methods for recovering clock signals that are multiples of symbol frequency, the PN 511 component of the DFS signal is match filtered to acquire symbol synchronization information and to synchronize counter circuitry that counts symbols, data segments, and data fields. The match filter response can provide approximate channel-impulse-response (CIR) from which initial equalizer coefficients can be calculated. These initial equalizer coefficients shorten the time it takes the known symbols in the DFS signal to complete the training of a Wiener type of adaptive equalizer.

In still other receiver designs, the VSB signal is equalized by decision-directed or blind equalization methods that do not rely on the special properties of the DFS signal. This was another of the ways that adaptive equalization can be done in the Grand Alliance prototype receiver design.

5.10.2 Specifications for the Data Field Sync

5.10.2.1 Data Segment Sync (DSS) Sequence

The same four-symbol Data Segment Sync sequence is used on all data segments, including Data Field Sync as well as each data segment containing payload data. The four successive symbols in the DSS sequence have normalized modulation levels of +5, –5, –5 and +5, respectively. The DSS sequences can be useful in equalization because they can independently establish the symbol clock phase and the times when data segments begin and conclude. Because the segment timing can be determined from the DSS sequences, simply identifying which segment contains the DFS signal is all that is necessary to establish when data fields begin and conclude. A full correlator is not required for detecting DFS signal, because the position of the sequence within the segment has already been established. A simple circuit that compares the data in each segment to the known PN511 sequence can accomplish this identification.

5.10.2.2 PN511 Sequence

The PN511 sequence immediately following the data segment sync in the initial data segment of each data field is a sequence of 511 symbols having normalized modulation levels each of which is either –5 or +5. These symbols each have the same approximately 93-nanosecond duration that each of the data symbols has.

5.10.2.3 Triple PN63 Sequences

The triple PN63 sequence immediately following the data segment sync in the initial data segment of each data field is a sequence of 189 symbols having normalized modulation levels each of which is either –5 or +5. These symbols each have the same approximately 93-nanosecond duration that each of the data symbols has. In alternate ones of the data fields all three of the successive PN63 sequences are identical. In the intervening data fields the polarity of the central PN63 sequences is inverted respective to the polarity of the preceding and succeeding PN63 sequences.

5.10.2.4 VSB Mode

Twenty-four two-level symbols identify the type of data in the data segments of the following data field.

5.10.2.5 Precode

This section of twelve symbols concluding the data segment containing the DFS signal repeats the last twelve symbols of payload data from the preceding Data Segment. This allows correction at the beginning of each data field of the symbols stored in the feedback loop of the 12-symbol NTSC-rejection comb filter, so that trellis decoding procedures halted during the DFS interval can resume without error.

5.11 Using Prescribed Symbols in the VSB Signal as Training Signal for Adaptive Equalization by a Wiener Method

5.11.1 Using the First 704 Symbols in DFS as Training Signal

Since the first 704 symbols in the DFS signal have prescribed values, they are suitable training signal for adaptive equalization by a Wiener method. These 704 symbols consist of the initial 4-symbol DSS sequence, the PN511 sequence, and the triple PN63 sequence.

5.11.2 Using the DSS Sequences as Training Signal

The initial 4-symbol DSS sequences in the 2nd through 313th segments of a data field comprise 1248 symbols of prescribed value, as compared to 704 symbols of prescribed value in the DFS signal. Accordingly, if adaptation of the equalization filtering uses a Wiener method, having the training signal include the DSS symbols as well as the symbols of prescribed value in the DFS signals will speed adaptation.

5.11.3 Using the Wiener Method to Augment the Kalman Method

Equalization error during the first 704 symbol epochs of the initial segment of each data field and during the first four symbol epochs of other data segments can be determined by comparing the equalized baseband DTV signal to the known values symbols should have at these times, as stored in read-only memory in the DTV receiver. Equalization error determined in this way is more likely to be correct than if determined by comparing the equalized baseband DTV signal to estimates of the transmitted signal based on minimal error from all possible symbol values.

So, a common practice is to intermix types of equalizer adaptation, using Wiener adaptation when the received symbols are known *a priori* by the receiver, and using Kalman adaptation when the symbols are unknown. Step size in the algorithm for incrementally adjusting the weighting coefficients of the equalizer can be somewhat larger during periods of Wiener adaptation than during periods of Kalman adaptation. This is because the confidence level as to the correctness of the estimates of originally transmitted symbols is higher when the received symbols have values known *a priori* by the receiver.

5.12 Initialization of Equalizer Weighting Coefficients Based on Channel Impulse Response (CIR) measurements

5.12.1 Computation of Initial Weighting Coefficients of the Adaptive Equalizer Based on CIR Measurements

The channel impulse response (CIR) can be measured based on the distortion of a signal with a uniform spectrum, using procedures as described in Sections 5.12.2 through 5.12.4, following. The CIR in the time domain can be normalized with respect to the principal signal component thereof. Complementing all the signal components of the normalized CIR except that or those associated with its principal signal component generates the time-domain impulse response of the channel-equalization filtering. The weighting coefficients of an IIR filter for suppressing longer-delayed post-echoes are corresponding terms of the time-domain impulse response of the channel-equalization filtering. The weighting coefficients of the feed-forward FIR filter, as is commonly used for suppressing pre-echoes and short-delay post-echoes, are approximated by corresponding terms of the time-domain impulse response of the channel-equalization filtering. The weighting coefficients of the feed-forward FIR filter are only approximately correct, however, because this technique does not take into account the fact that the feed-forward FIR filter attempts to cancel echoes with variously delayed response to its input signal. This input signal is not-echo free and its echoes give rise to repeat echoes. These repeat echoes have negligible energy only if the principal signal of the feed-forward-FIR-filter input signal is much larger than its accompanying echoes. This simple method of extracting the weighting coefficients of the feed-forward FIR filter from the normalized CIR does not take the repeat echoes into account.

The Dietrich-Greenberg method for computing initial weighting coefficients for an all-FIR equalization filter takes repeat echoes into account.[21] The normalized CIR is convolved with itself to generate a synthetic normalized CIR. Complementing all the signal components of the synthetic normalized CIR except that or those associated with its principal signal component generates the time-domain impulse response of the channel-equalization filtering, taking repeat echoes into account. Convolution can be done in the time domain. Alternatively, convolution can be done in the DFT domain, multiplying the DFT of the reception channel term-by-term by itself; and the I-DFT of the DFT formed from the resulting product terms provides the synthetic normalized CIR.

When a decision-feedback equalizer having a feed-forward FIR filter and an IIR filter including an FIR feedback filter is used, J. D. McDonald has modified the Dietrich-Greenberg method in the following way. The weighting coefficients for the IIR filtering are determined by the direct method outlined in the second paragraph previous, since the decision-feedback terms are presumably echo-free. Only the terms of the normalized CIR that are not cancelled by the IIR filter are convolved with the normalized CIR to synthesize the synthetic normalized CIR that the feed-forward FIR filter must provide equalization for. Complementing all the signal components of the synthetic normalized CIR except that or those associated with its principal signal component generates the time-domain impulse response of the feed-forward FIR filter, taking repeat echoes into account.

5.12.2 Determination of CIR from Equalizer Training Signal

An equalizer training signal as received over-the-air can be de-convolved by the equalizer training signal as known *a priori* at the receiver to determine the CIR of the reception channel,

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[21] C. B. Dietrich and A. Greenberg: "Deghosting Apparatus Using Pseudorandom Sequences", U. S. patent No. 5 065 242, 23 August 1994.

presuming that the equalizer training signal has a uniform spectral response. This de-convolution is best implemented by DFT methods. The DFT of the equalizer training signal received over the air is divided term-by-term by the DFT of the equalizer training signal as known *a priori* at the receiver, to determine the DFT of the reception channel. The term-by-term division is usually done using log and anti-log look-up tables in ROM. The inverse-transform (I-DFT of the DFT of the reception channel is the CIR in the time domain.

The PN511 sequence in the DFS signal is a single cycle of a maximal-length PN sequence. Each PN63 sequence in the DFS signal is maximal-length, also. A maximal-length PN sequence has special properties as an equalizer training signal, since the de-convolution process can be supplanted by match filtering using the PN sequence as kernel. This implements a correlation procedure that generates the CIR as a portion of the match filter response. This works because the auto-correlation function of a PN sequence exhibits a peak response at a specific phasing of the PN sequence that is much higher than the response at other phasings. Developing the CIR by correlation tends to introduce less noise than developing the CIR by de-convolution. This is because developing the CIR by correlation avoids division, DFT and I-DFT steps that introduce rounding errors and binning errors.

5.12.3 Re-Computation of CIR and Equalizer Coefficients using a Wiener Method on Recycled Training Signal

A block of consecutive DTV symbols as originally received comprising the continuous equalizer training sequence together with surrounding data can be stored by the receiver. After the weighting coefficients of the equalizer have been initially determined, this block of consecutive DTV symbols is passed through the equalizer so the initial weighting coefficients can be refined by a Wiener method applied to known symbols in the equalizer response. The block can be passed through the equalizer once or, in an alternative design, may be looped around to pass through the equalizer several times.[22] E.g., the block of data consecutive DTV symbols can be three data segments long, consisting of the final data segment of a data field and the first two segments of the next data field. Operating at normal clock rates, this block can be re-cycled through the equalizer as many as a hundred times or so during each data field, for refining the weighting coefficients of the equalizer.

5.12.4 Determination of CIR from Long Data Sequences

The symbols in a baseband DTV signal occur quite randomly. A symbol sequence a few thousand symbols long can be selected from a longer symbol sequence that is de-convolved using the selected symbol sequence, thus to determine the DFT of the reception channel. This de-convolution is best implemented by DFT methods, dividing the DFT of the longer symbol sequence received over the air term-by-term by the DFT of the trellis decoding decisions, to determine the DFT of the reception channel.[21] The term-by-term division is usually done using log and anti-log look-up tables in ROM. The I-DFT of the DFT of the reception channel is the CIR in the time domain.

Because of the duration of the selected symbol sequence being many times longer than the echo range, the edge effects of echoes of adjoining data overlapping the selected symbol sequence are minimized in the DFT of the reception channel. The edge effects generally have considerably less effect on the amplitudes of the complex time-domain CIR components than is the case for such components as determined from the equalizer training signal.

[22] O.Piirainen: "Method for Estimating Impulse Response, and Receiver", U. S. patent No. 6 327 315, 4 December 2001.

The amplitudes of the complex time-domain CIR components strongly tend to be accurate, but accuracy of the phases of these components is apt to be less accurate under conditions of dynamic multipath reception. This is because the processing delay in determining the CIR is appreciably long in this method. The phases of the complex time-domain CIR components can be corrected by temporal compensation procedures based on DFT properties.

Alternatively, the phases of the complex time-domain CIR components can be modified to correspond with the phases of corresponding complex time-domain CIR components as determined from the equalizer training signal. The determination of CIR from long data sequences by de-convolution methods is usually used to correct edge effects in the auto-correlation procedures used to measure CIR from PN sequences in the DFS signal. Components appearing in the CIR extracted by PN511 sequence auto-correlation procedures that do not appear in the CIR extracted by de-convolution of a long data sequence are presumed to be attributable to edge effects and are expunged.

5.13 Speed of Operation in the Adaptive Equalization Filtering of the DTV Receiver

If multipath reception conditions are changing, it is important that the adaptive equalization filtering in the DTV receiver be able to initialize its weighting coefficients reasonably accurately based on a time interval containing no more than a few hundred symbols. Reasonably fast and accurate initialization based on CIR measurements is most easily done using the time interval containing the first 704 symbols of prescribed value that begin each data field. The fact that the first 704 symbols of prescribed value that begin each data field are available only about every 24 milliseconds limits the speed of initialization based on CIR measurements. However, initialization based on CIR measurements can be done in less than a frame time, while initialization based on a Weiner method can take several frame times. Alternatively, initialization of equalizer coefficients can be based on decision-directed CMA; see Section 6.13.1.6 ("Initialization of Adaptive Equalizer Coefficients using Constant Modulus Algorithm (CMA)"), or on MRCA see Section 6.13.1.7 ("Initialization of Adaptive Equalizer Coefficients using Modified Reduced Constellation Algorithm (MRCA)").

After the equalizer weighting coefficients are initialized to reasonably correct values, a type of feedback called auto-regression can be employed to adapt the equalizer weighting coefficients so as to converge those coefficients to values that track equalization reasonably closely to changing multipath reception conditions. Auto-regressive adaptation of equalizer weighting coefficients has a characteristic common to feedback loops that the hold-in to a tracking condition is easier to maintain than the pull-in from a non-tracking condition to a tracking condition.

Speed of convergence to correct equalizer parameters is not the only important criterion of performance of the auto-regression algorithm. Ultimate accuracy and response to noise are also important. Adaptive equalization methods that use auto-regression generally proceed by successive incremental adjustments of equalizer coefficients towards desired values for them and such methods typically exhibit convergence times of several data segment intervals unless initialization is quite accurate.

By altering the rate of incremental adjustment of the equalizer coefficients, speed of convergence can be traded off for accuracy and noise immunity. The success of such a trade-off may depend on non-linear techniques; for example, switching between a quicker-converging acquisition mode and a slower-converging refinement mode, or using varying step sizes in a steepest-descent technique. Techniques for affecting the rate of convergence and its final accuracy are treated in the literature.

The kernel width of an adaptive equalizer determines the maximum permissible step size for the LMS (estimated-gradient) algorithm generally used for adapting the weighting coefficients, however. The maximum permissible step size is reduced by a factor equal to the number of kernel taps, from the maximum step size permitted in the steepest-descent (known gradient) algorithm.[23] The reduction in step size accommodates the need to average out the random fluctuations of the updates to the weighting coefficients caused by noise and the randomness of the data pattern. The equalizers with about 50 microsecond echo-suppression range that were the general rule in the 1995 to 2000 timeframe converge almost twice as fast using the LMS algorithm as equalizers with a 90-microsecond echo-suppression range would.

5.14 Symbol-Sequential vs. Block-Sequential Decision-Directed Adaptation of Equalizer Coefficients

In symbol-sequential adaptation, at least one weighting coefficient in the equalizer is updated every symbol interval. Presuming that the input signal to the equalizer is in normal time order, the change in equalizer coefficients should never be retrograde in time.[24] An equalizer that employs the LMS algorithm converges substantially faster if the weighting coefficients in the equalizer are all updated concurrently, rather than being updated sequentially.[25] [26] Such symbol-sequential adaptation can be done on a pipeline basis, without the latent delay suffered in adaptation procedures that rely on time reversal.[27] Symbol-sequential adaptation that simultaneously updates all the weighting coefficients in the equalizer every symbol interval is possible when the equalizer is implemented in monolithic integrated circuitry. Owing to hardware limitations, smaller blocks of taps were updated on successive symbol intervals in the Grand Alliance prototype receiver.

In symbol-sequential adaptation that employs the LMS algorithm, the gain factor μ that controls incremental changes of the weighting coefficients often is chosen quite small. This is so that occasional impulse noise is less likely to change the weighting coefficients in the equalizer appreciably and generate equalizer errors serious enough to cause the adaptation algorithm to lose tracking. The effects of such occasional noise bursts can be reduced by rejecting error signals generated during impulse noise occurrences.[28] This allows μ to be made somewhat larger without too much risk of loss of tracking by the adaptation algorithm.

In block-sequential adaptation all equalizer weighting coefficients are updated simultaneously at prescribed intervals, rather than being updated every symbol interval, and every symbol is equalized by weighting coefficients from the same block. Since the incremental changes of the weighting coefficients are accumulated before their application, the applied weighting coefficients are considerably less sensitive to noise in the equalizer response. Occasional noise bursts are less likely to change the weighting coefficients applied to the equalizer sufficiently to cause the adaptation algorithm to lose tracking. Consequently, a larger

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[23] R. D. Gitlin, J. F. Hayes, and S. B. Weinstein: *Data Communications Principles*, Plenum Press, N.Y. and London, pg. 551, 1992.

[24] T. G. Laud: "FIR Filter Coefficient Updating System", U. S. patent No. 5 392 315, issued 21 February 1995.

[25] W. H. Paik, S. A. Lery, and A. Wu: "Method and Apparatus for Updating Coefficients in a Complex Adaptive Equalizer", U. S. patent No. 5 243 624, 7 September 1993.

[26] B. J. Currivan and J. E. Ohlson: "Dynamically Adaptive Equalizer System and Method", U. S. patent No. 5 416 799, 16 May 1995.

[27] A. L. R. Limberg: "Dynamically Adaptive Equalizer System and Method", U. S. patent No. 5 901 175, 4 May 1999.

[28] L. E. Nielsen: "Adaptive Equalizer with Impulse Noise Protection", U. S. patent No. 5 692 010, 25 November 1997.

gain factor μ can be safely chosen, which tends to speed up convergence. Because of the larger number of digital multiplications that must be performed each symbol interval, symbol-sequential adaptation that simultaneously updates all the weighting coefficients in the equalizer every symbol interval consumes more power than block-sequential equalization. Block-sequential equalization has been reported to provide faster convergence and better tracking than symbol-sequential equalization when LMS algorithms are employed.[19] There is a pipeline approach to block-sequential equalization that avoids the need for time-reversal procedures.[29]

A question that arises when contemplating the use of block-sequential adaptation is how frequently to update the equalizer coefficients. Simulations have shown that updating at 2048-symbol-epoch intervals, cutting back to 512-symbol-epoch intervals when extreme multipath conditions are encountered, is satisfactory.[30]

5.15 Time-Domain Aliasing Problems in FIR Filtering of DTV Signal that has a Weak-Energy Principal Component

Pre-echoes and at least short-delayed post-echoes are suppressed by an FIR portion of the equalization and echo-suppression filtering. Suppression is achieved by combining variously delayed responses to the equalizer input signal. The equalizer input signal contains echoes which remain as time-displaced repeats of echoes after each original echo is suppressed by being differentially combined with the principal component of a suitably attenuated and delayed response to the equalizer input signal. The repeats of echoes have less energy than the original echoes only to the extent that the principal component of the equalizer input signal has significantly more energy than any of the echoes that are to be suppressed by the FIR portion of the equalization and echo-suppression filtering.

One strategy for dealing with no component of the equalizer input signal having significantly more energy than other echoes is to attempt to use the earliest of the components with more energy as principal component. An IIR portion of the equalization and echo-suppression filtering can suppress most later components without generating any repeats of them, because the echo-free response of the filtering is used for suppressing all post-echoes except short-delay ones in the IIR portion of this filtering. Short-delay post-echoes are too close in time to the principal component to be suppressed by IIR portions of the equalization and echo-suppression filtering, because loop delay in this recursive filtering prevents their suppression. A short-delay post-echo or a pre-echo that has substantially as much energy as the principal component will be difficult to suppress in the FIR portions of the equalization and echo-suppression filtering, since the energy of repeated echoes dwindles in energy to insignificant levels only after several cycles of repetition. Spectrum combining techniques can alleviate this sort of problem.

5.16 Spectrum-Combining Techniques

Sometimes multipath reception conditions are such that none of the reception paths generates a component response having substantially more energy than any of the one or more component responses to other reception paths. In such an instance, combining the baseband DTV signal with itself, as delayed by the interval between the earliest two of the stronger component responses, can provide channel-equalization and echo-suppression filtering an input signal with a principal component response that is likely to have substantially more energy than any echo. This

[29] Jian Yang, C. B. Patel, Tianmin Liu, and A. L. R. Limberg: "Rapid-Update Adaptive Channel-Equalization Filtering for Digital Radio Receivers, such as HDTV Receivers", U. S. patent No. 5 648 987, 15 July 1997.

[30] C. B. Patel, A. L. R. Limberg, and J. D. McDonald: PLM Presentation Before ATSC T3/S9, Washington D.C., 7 June 2001 (slide 6).

combining can be done in a pre-filter that precedes conventional channel-equalization and echo-suppression filtering. The CIR, as determined from the training signal, is used as the basis for choosing the delay to be used in the pre-filter and for choosing whether the combining is to be additive or differential. The combining procedure results in additional pre-echoes and post-echoes. However, it generates a principal component of reception that generally has significantly more energy than any other component. So, in the FIR portions of the equalization and echo-suppression filtering, repeats of pre-echoes and short-delay post-echoes can be reduced to negligible energy within fewer cycles of repetition.

Another spectrum-combining technique is to determine the CIR and then pre-filter the baseband DTV signal with a time-domain filter the coefficients of which mirror the CIR in time.[31] The time-domain impulse response of the pre-filter is a principal component of reception flanked by an ensemble of pre-echoes and an ensemble of post-echoes. These ensembles have mirror symmetry around the principal component. The pre-filter response is then supplied for conventional channel-equalization and echo-suppression filtering.

5.17 Sparse Equalization

Each adaptable weighting coefficient in an adaptive equalizer introduces an extra mean-squared-error term owing to the stochastic jitter associated with the adaptation process, which extra MSE term is referred to as *excess MSE* (EMSE). If echo conditions remain substantially the same over time, the amount of MSE is proportional to the number of weighting coefficients in the equalizer and the step-size μ used for adaptation. While a larger step-size μ may be desirable for faster equalizer tracking, it induces more EMSE. Similarly, although more weighting coefficients in the equalizer can minimize steady-state MSE, more than 500 taps at symbol-epoch spacing have been used in equalizers in DTVs for the consumer market. This severely constrains the step-size μ, so EMSE can be kept small enough that threshold of visibility (TOV) can be obtained at reasonably low SNR. Small step-size μ will slow or even stall blind equalization, and it slows equalizer tracking during rapid fades.

In sparse equalization small-valued weighting coefficients are forced to be zero-valued. Accordingly, there are apt to be different intervals between adjacent ones of the non-zero weighting coefficients that are left. Sparse equalization reduces the number of multiplication operations used for equalization and reduces EMSE.[9] The reduction in EMSE allows the step-size μ to be increased, allowing faster equalizer tracking during rapid fades. However, the low-strength echoes that are not canceled when sparse equalization is employed combine to form a type of "noise" that raises the SNR required at TOV.

A fundamental problem in the design of sparse equalizers is determining which weighting coefficients are to be forced to zero value. An ad hoc method for optimally choosing the weighting coefficients to be forced to zero value takes advantage of the observation the MSE and CM cost functions exhibit sparse minima settings that are in close proximity.[32] [33] In equalizers in which the CIR is periodically determined from the training signal, the locations of strongest echo

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[31] R. Citta, X. Wang, Y. Wu, B. Ledoux, S. Laflèche, and B. Caron: "A VSB DTV Receiver Designed for Indoor and Distributed Transmission Environments", IEEE 52nd Annual Broadcast Symposium, Washington, D.C., October 9–11, 2002.

[32] T. J. Endres, R. A. Casas, S. N. Hulyalkar, and C. H. Strolle: "On Sparse Equalization Using Mean-square-error and Constant Modulus Criterion", Conference on Information Sciences and Systems, Princeton, N.J., pp. TA7b:7–12, March 2000.

[33] T. J. Endres, S. Hulyalkar, T. A. Schaffer, and C. H. Strolle: "Reduced Complexity Equalizer for Multi Mode Signaling", U. S. patent No. 6 426 972, 30 July 2002.

components can be used as a basis for determining which weighting coefficients should not be forced to zero value.

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John G. Proakis: *Digital Communications*, second edition, McGraw-Hill, New York, N.Y., 1989.

M. H. Lee: "Ghost Canceling Method and Apparatus Using Canonical Signed Digit Codes", U. S. patent No. 5 623 318, 22 April 1997.

S. Haykin: *Adaptive Filter Theory*, Prentice Hall, 1991.

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P. Fabre and C. Gueguen: "Improvement of the Fast Recursive Least-Squares Algorithms via Normalization: A Comparative Study", *IEEE Trans. Acoustics, Speech & Signal Processing*, vol ASSP-34, pp. 296–308, April 1986.

S. Ljung and L. ljung: "Error Propagation Properties of RLS Adaptation Algorithms", *Automatica*, March 1985.

6. DESIGN TECHNIQUES FOR DTV RECEIVERS

This section describes design techniques for portions of digital television signal receivers that were not included in the Grand Alliance prototype receiver used for system testing [3]. Some of these designs have already been embodied in commercially available receivers; others may be embodied within receivers in the near future.

Some of these design techniques are alternatives to each other and cannot be used together in the same DTV receiver.

6.1 RF Signal Acquisition

As of the end of 2002, carriage of ATSC DTV signals on cable systems was not required, so reception of these signals required a suitable broadcast antenna. For many years, the widespread use of cable to receive NTSC signals, including local broadcasts as required by "must carry" rules, has reduced the need for and the development of such antennas. Opportunities exist for the application of recent technologies to antenna reception of DTV broadcast signals. Many reception problems can be mitigated by use of a mast-mounted low-noise amplifier (LNA). Currently, several manufacturers sell LNAs.

EIA/CEA-909 "Antenna Control Interface" concerns improving the performance of outdoor, attic, and indoor antennas by allowing the DTV (or analog NTSC) receiver to control the antenna direction, LNA gain, polarization, and other parameters on a channel-by-channel basis. This facilitates optimal adjustments of reception to be made automatically, without a user having to make separate individual adjustments when different stations are successively selected for reception.

Channel impulse response measurements can be used for generating an on-screen display to aid in antenna alignment.[34] Alternatively, equalizer weighting coefficients can be used for generating an on-screen display to aid in antenna alignment.[35]

6.2 Tuner

6.2.1 RF Amplifier

6.2.2 Wideband Linear Mixer for Analog RF Signal

6.2.2.1 Up-Converting First Mixer

A commutating type single-balanced resistive RF mixer with good intermodulation characteristics for UHF IF signal even when RF input signal is high-level includes a bifilar balun transformer coupling the RF input signal to the input winding of a trifilar balun transformer. The two output windings of the trifilar balun transformer are connected in series with the drain-source paths of a pair of GaAs FET transistors having grounded sources. A local oscillator has a symmetric output signal that is coupled to a self-bias network and also to the gate electrodes of those FETs. An IF output signal is taken between ground and a common junction of the output balun windings.[36] There is also a double-balanced form of this mixer.[37] These arrangements provide high-level mixers with intermodulation characteristics for ultra high frequencies that allegedly are better than prior art commutating mixers employing diodes.

6.2.2.2 Down-Converting First Mixer

A wide-range linear mixer that down-converts to 44 MHz center frequency can exhibit a 2 dB or so better noise figure than one that up-converts to 920 MHz center frequency. Up-converting to 920 MHz center frequency before down-converting to 44 MHz center frequency simplifies image-frequency rejection. However, RF signals can be mixed with complex local oscillations and the resulting signals combined in a phase shift network that generates a down-converted IF signal in which image frequencies are suppressed.[38]

A wide-range linear mixer can be constructed that drives an ensuing VHF IF amplifier from a selected one of a plurality of paralleled mixers that degenerate the RF input signal with different amounts of linearizing feedback. Mixer selection is done using a zonal ("gated") AGC technique.[39]

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34 R. W. Citta and G. J. Sgrignoli: "Antenna Orientation System for Digital TV Receiver", U. S. patent No. 5 574 509, 12 November 1996.

35 R. Iwamura: "System and Method for Aligning an Antenna", U. S. patent No. 5 940 028, 17 August 1999.

36 P. Dobrovolny: "High Level Wide Band RF Mixer", U. S. patent No. 5 027 163, 25 June 1991.

37 P. Dobrovolny: "Double-balanced High Level Wideband Mixer", U. S. patent No. 5 280 648, 18 January 1994.

38 H. Guegnaud and M. Robbe: "Mixer Device with Image Frequency Rejection", U. S. patent No. 5 901 349, 4 May 1999.

39 K.W. Clayton and R. R. Rotzoll: "System and Method for Switching an RF Signal between Mixers", U. S. patent No. 5 805,988, 8 September 1998.

6.2.3 Local Oscillator

A plurality of local oscillators may be used for up-converting from different television broadcast bands to a UHF intermediate frequency in a dual-conversion receiver.[40]

6.2.4 Frequency Synthesizer Considerations

Phase noise is critical to digital modulation systems as opposed to analog. For ATSC 8-VSB, the threshold is –77 dBc at 20 KHz offset. Therefore, the local oscillator system should be significantly better to not deteriorate system performance.

During the transition period, NTSC or VSB DTV signals are apt to be received using the same RF amplifier and first converter. The operation of the frequency synthesizer typically is controlled by a microprocessor that includes a time-out arrangement that starts when a channel is initially selected for reception. The microprocessor instructs the frequency synthesizer to generate the expected carrier frequency for a VSB DTV signal when a channel is initially selected for reception. If lock to the pilot carrier of a VSB DTV signal is achieved, no further action is taken until another channel is selected. If lock to the pilot carrier of a VSB DTV signal is not achieved within a predetermined time after a channel is initially selected for reception, the time-out arrangement in the microprocessor instructs the frequency synthesizer to generate the expected carrier frequency for an NTSC signal.[41]

6.2.5 Automatic Fine Tuning (AFT)

DTV receivers that use IF filtering to establish square-root-raised-cosine roll-offs of the reception channel frequency response are very sensitive to mis-tuning. These DTV receivers usually employ AFT so that the received DTV signal is filtered properly. AFT may be applied to the offset oscillator in a frequency synthesizer used for tuning. In a plural-conversion receiver AFT may instead be applied to the second local oscillator.

In some designs, the bandpass filter used for extracting the frequencies near DTV carrier frequency for application to the AFT detector has a tilt in its amplitude response that compensates for the roll-off of the DTV signal.[42] If the carrier-side roll-off of the final-IF DTV signal is established by digital filtering, the amplitudes of pilot carrier in the response of the roll-off filter and the response of a complementary filter can be compared to develop an AFT signal that aligns the receiver for correct carrier-side roll-off. The Nyquist roll-off can then be established by digital filtering of baseband DTV signal to complete the design of an automatically aligned DTV receiver.

An AFT signal is developed quite readily in a DTV receiver that establishes square-root-raised-cosine roll-offs by SAW filtering the final IF signal and that subsequently synchrodynes the final IF signal to baseband without intervening frequency conversion. AFT is applied to an AFT's local oscillator preceding the final IF signal amplifiers and the SAW filtering that defines the bandwidth of the reception channel. Another local oscillator used for down-converting the final IF signal to baseband is a crystal-controlled oscillator oscillating at a fixed frequency that is the same as the carrier frequency of a DTV signal would be if it were properly aligned with the SAW filter frequency response. The synchrodyne is performed using in-phase and quadrature

40 P. Dobrovolny: "Bandswitched Tuning System Having a Plurality of Local Oscillators for a Digital Television Receiver", U. S. patent No. 5 420 646, 30 May 1995.

41 R. W. Citta: "Dual HDTV/NTSC Receiver Using Sequentially Synthesized HDTV and NTSC Co-channel Carrier Frequencies", U. S. patent No. 5 283,653, 1 February 1994.

42 A. L. R. Limberg: "Automatic Fine Tuning of Receiver for Digital Television Signals", U.S. patent No. 6 445 425, 3 September 2002.

synchronous detectors. AFT signal for controlling the frequency of the AFT's local oscillator can be generated by lowpass filtering the baseband response of the quadrature synchronous detector.

Alternatively, AFT signal can be developed by the Costas method. In the Costas method the baseband response of the in-phase synchronous detector is filtered to suppress zero-frequency component and then used to multiply the baseband response of the quadrature synchronous detector. Lowpass filtering of the resulting product signal generates the AFT signal. The use of the Costas method is advantageous in that there is less likelihood of failure of AFT loop pull-in when the pilot carrier is attenuated by multipath or by being too far down on the slope of the IF passband filtering for some reason.

Until phase lock is established, the bandwidth of the lowpass filtering of the AFT detector may be broadened to increase the pull-in range of the AFT.[43] When phase lock is established, the bandwidth of the lowpass filtering of the AFT detector is then narrowed to reduce phase noise in the first LO. Carrier offsets can move the pilot carrier outside the range of a narrowband AFT detector. The receiver can be constructed to include circuitry that infers this to be the reception condition if carrier phase lock is not achieved after a prescribed time, but there is other indication that a VSB DTV signal is being received. The receiver can be constructed to include further circuitry for adjusting the frequency of a local oscillator in discrete steps until the narrowband AFT detector can establish carrier phase lock. The number of discrete steps needed for carrier phase lock to be established can be stored in channel memory to reduce the time for achieving lock the next time the channel is tuned to.[44] [45] Alternatively, information concerning the amount of carrier offset can be stored in more direct form in the channel memory.[46]

In some receiver designs, the gain before the AFT detector is boosted until carrier phase lock is established.[47]

6.2.6 Bibliography of Documents Regarding RF Amplifier and First Converter but not Specifically Cited

V. Birleson: “Interference-free Broadband Television Tuner”, U. S. patent No. 5 847 612, 8 December 1998.

6.3 Gain Control in the Front End of the Receiver

The front end of a DTV receiver designed for the consumer market comprises RF amplifier, mixer and IF amplifier stages for selecting an RF DTV signal, converting it in frequency to IF DTV signal(s), and amplifying the IF DTV signal(s). These procedures are carried out in the analog regime and should maintain linearity of the DTV signal over as large a dynamic range as possible. Linearity should be maintained over the entire dynamic range of the DTV signal and any accompanying random noise that is not so large as to obliterate modulation levels most of the time. Linearity should be maintained over the entire dynamic range of the DTV signal and

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43 G. Krishnamurthy, V. G. Mycynek, and G. J. Sgrignoli: “Carrier Acquisition by Applying Substitute Pilot to a Synchronous Demodulator During a Start Up Interval”, U. S. patent No. 5 410 368, 25 April 1995.

44 D. S. Han: “Digital Carrier Wave Restoring Device and Method for Use in a Television Signal Receiver”, U. S. patent No. 5 809 088, 15 September 1998.

45 D. S. Han: “Method for Reducing Carrier Recovery Time in High Definition Television Receiver”, U. S. patent No. 5 933 200, 3 August 1999.

46 J. K. Lee: “Apparatus for Finely Adjusting Tuning Data for a Television Receiver and the Method thereof”, U. S. patent No. 5 428 405, 27 June 1995.

47 G. Krishnamurthy, V. G. Mycynek, and G. J. Sgrignoli: “AGC System with Overriding Maximum Gain During an Initial Interval to Enhance Signal Acquisition”, U. S. patent No. 5 546 138, 13 August 1996.

any co-channel NTSC interference the demodulation artifacts of which are to be suppressed by selective filtering.

6.3.1 Avoidance of Forward Automatic Gain Control in DTV Designs

Forward AGC, in which a device is at times operated at saturation current levels in order to reduce transconductance, is used for gain control of analog TV signals in an RF-amplifier, frequency-conversion, or early-IF-amplifier stage because gain reduction does not decrease SNR significantly. The non-linearity of forward gain control is tolerated by signals like the gamma-corrected analog TV signal, the modulation of which tends to be logarithmic in character. The non-linearity of forward gain control is a problem for 8-VSB DTV signals or any other PAM signals with multiple modulation levels that are to be maintained evenly spaced.

6.3.2 Reverse Automatic Gain Control in DTV Designs

Reverse AGC, in which a device is at times operated with reduced current levels in order to reduce transconductance, is used for gain control of analog TV signals in later IF amplifier stages. Gain control in earlier stages limits the dynamic range of the input signal to the later IF amplifier stage so that it operates with acceptable linearity. Reverse AGC can be used for a later IF amplifier stage in a DTV receiver. However, in order to have acceptable linearity in the amplified 8VSB signal, the dynamic range of the input signal to the later IF amplifier stage has to be constrained more than it usually is in an analog TV receiver design.

6.3.3 Independent AGC of RF Amplifier to Avoid Cross-Modulation in Mixer

AGC signals derived from IF signals that have been filtered to 5.38 to 6 MHz bandwidth are unresponsive to adjacent-channel energy. Controlling the gain of the RF amplifier by delayed application of such AGC signal does not reduce the gain sufficiently to avoid overloading the first mixer when there is a strong adjacent-channel signal. This results in non-linearity in the first mixer and cross-modulation of the strong adjacent-channel signal with the desired-channel signal. While these effects can be tolerated to some degree in an analog TV receiver, they should be avoided in a receiver of multi-level pulse-amplitude-modulation signal. This mixer overload problem can be avoided by amplifying the first mixer output signal in a wideband IF amplifier and deriving AGC for the RF amplifier from the response of the wideband IF amplifier.

Delayed AGC is not used when there is completely independent AGC of the RF amplifier. However, delayed AGC that utilizes the underlying principle of independent AGC of the RF amplifier is known.[48]

Plural-loop AGC is often used in digital DTV receiver designs.[49]

6.3.4 Variable Coupling Network for Gain Control

6.3.4.1 Variable Coupling Network for RF Signal

CATV tuners use PIN-diode variable coupling networks for RF gain control.[50] This approach can be adapted for RF gain control in tuners for over-the-air signals.

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[48] Y. Igarishi, H. Mizukami, and T. Nagashima: “High-Definition Television Signal Receiving Apparatus And Gain Control Circuit Thereof”, U. S. patent No. 5 940143, 17 August 1999.

[49] E. Nakanishi and T. Onodera: “Linear Receiver having Dual Automatic Gain Control Loops”, U. S. patent No. 5 722 062, 24 February 1998.

[50] S. Masuda: “AGC Apparatus Concurrently Satisfying Sufficient Impedance Matching Characteristic and Linear AGC Characteristic”. U. S. patent No. 5 737 033, 7 April 1998.

6.3.4.2 Variable Coupling Network for UHF IF Signal

A variable coupling network between a pair of receiver stages, such as the first mixer and the first IF amplifier, can provide gain control with acceptable linearity for maintaining even spacing between multiple modulation levels of 8-VSB DTV signals. In one design of a dual-conversion receiver with a 920 MHz IF, the variable coupling network comprises a quadrature coupler with two of its four ports terminated by respective series connections, each of a resistor in series with a PIN diode. The other two ports of the quadrature coupler are an input port connected for receiving the output signal from the first mixer and an output port connected for supplying input signal to the first IF amplifier. An AGC voltage controls the conductances of the PIN diodes for adjusting gain through the quadrature coupler.[51] The quadrature coupler is designed for a specific frequency. So, this variable coupling network is impractical for controlling RF amplifier gain, presuming the RF amplifier to be tunable for selecting any of a number of DTV signals for reception. The quadrature coupler is constructed within a quarter-wavelength waveguide, so its dimensions become too large to be practical at VHF IF.

6.3.4.3 Variable Coupling Network for VHF IF Signal

Variable resistive coupling to an emitter-coupled differential-amplifier connection of bipolar transistors can employ semiconductor diodes as shunt elements in resistive voltage division.[52] The distortion in the voltages across the semiconductor diodes is compensated for by the non-linearity of the transconductances of the emitter-coupled bipolar transistors.

6.3.5 Controlled-Gain Amplifiers

Controlled-gain amplifiers developed for analog color television receivers usually have sufficient linearity for DTV signals, presuming they are not driven into an overload condition.

6.3.5.1 Controlled-Gain Amplifiers Employing Current-Splitter Techniques

A gain-controlled later-IF amplifier stage for VHF DTV signal can comprise a common-emitter (or common-source) amplifier transistor and a pair of current-splitting bipolar transistors connected for passing the collector or drain current of the amplifier transistor through an interconnection of their emitter electrodes. AGC voltage is applied between the bases of the current-splitting bipolar transistors for controlling the gains of their respective cascode connections with the amplifier. One of these cascode connections is in the forward path for the VHF IF signal. The other of these cascode connections can be in a feedback path for the VHF IF signal that degenerates the input signal supplied to the amplifier transistor.

Gain control by current splitting does not affect amplifier linearity. The additional gain control afforded by degenerating the input signal supplied to the amplifier transistor permits an earlier IF amplifier to be operated at higher gain, tending to improve SNR when Johnson noise from that earlier IF amplifier is a significant component of noise.

6.3.5.2 Controlled-Gain Amplifiers Employing Controlled Degeneration

Four diodes can be connected in full-bridge connection for degenerating the emitter-to-emitter connection of an emitter-coupled differential-amplifier connection of bipolar transistors. The emitter degeneration of the emitter-coupled differential-amplifier bipolar transistors is controlled

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[51] P. Dobrovolny: "AGC Circuit For Double Conversion Digital Television Tuner", U. S. patent No. 5 270 824, 14 December 1993.

[52] M. J. Gay: "Integrated Circuit Amplifier having Controlled Gain and Stable Quiescent Output Voltage Level", U. S. patent No. 3 962 650, 8 June 1976.

by varying the current flowing through the 4-diode bridge.[53] Alternatively, the emitter-to-emitter degeneration of an emitter-coupled differential-amplifier connection of bipolar transistors can be provided by the paralleled source-to-drain paths of insulated-gate field transistors operated in controlled saturation responsive to voltages applied to their gate electrodes.[54]

6.3.5.3 Controlled-Gain Amplifiers Employing Controlled Ratio of Load and Degeneration Resistances

Field-effect-transistor switches select one of various load resistors and one of various degeneration resistors for an amplifier transistor in some designs for controlled-gain amplification.[55] [56]

6.3.6 Analog AGC

The coherent AGC described in [3] for the prototype Grand Alliance receiver is a slow-acting AGC with time constant of several hundred microseconds, similar to that used in analog TV receivers. Faster-acting AGC has been found preferable in DTV, particularly when dynamic multipath reception conditions are encountered.

6.3.7 AGC Developed from Average Amplitude of Data

If the received signal is a trellis-coded 8-VSB signal, except for the initial data segments of each field and the DSS sequences, all eight of its modulation levels are equally likely to occur. Accordingly, peak signal excursions can be inferred from average detection of the rectified baseband DTV signal, providing a basis for developing AGC signal. The presence of the pilot carrier is compensated for in this average detection process.[57] DSS sequences and the DFS signal are preferably provided special treatment in the average detection procedure to account for their absolute amplitudes departing somewhat from average absolute amplitude.[58] [59] If demodulation is performed in the analog regime, this AGC method protects the dynamic range of the A/D converter that follows. This AGC method is reasonably insensitive to occasional impulse noise; and that insensitivity can be improved by inverting high-energy impulse noise to previously detected average signal levels. If the received signal is a mixture of trellis-coded 8VSB signal and another form of VSB signal, it is apt to be more difficult to infer peak signal excursions from average signal level.

More sophisticated forms of AGC based on average energy of data may respond to the average amplitude of the vector sum of the orthogonal-phase (e.g., in-phase and quadrature-phase) baseband signals, so AGC does not fluctuate when complex demodulation is not done at correct phasing. The orthogonal-phase baseband signals have their direct terms suppressed before being used to address read-only memory (ROM) storing look-up tables (LUTs) for squares. The squares are summed. Square-rooting the sum of the squares using further ROM storing LUTs for square roots will generate the vector sum of the orthogonal-phase baseband

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53 L. Gotz and H. Kowatsch: “High Frequency Amplifier”, U. S. patent No. 4 531 097, 23 July 1985.

54 T. L. Viswanathan: “Programmable Gain Amplifier” U. S. patent No. 6 018 269, 25 January 2000.

55 R. W. Ezell: “Highly Linear Variable-Gain Low Noise Amplifier”, U. S. patent No. 6 100 761, 8 August 2000.

56 R. W. Ezell: “Method for a Highly Linear Variable-Gain Low Noise Amplifier”, U. S. patent No. 6 218 899, 17 April 2001.

57 R. W. Citta, D. M. Mutzabaugh, and G. J. Sgrignoli: “AGC System with Pilot Using Digital Data Reference”, U. S. patent No. 5 565 932, 15 October 1996.

58 G. Krishnamurthy and R. Turner: “Sync Compensated AGC System for VSB Receiver”, U. S. patent No. 5 764 309, 9 June 1998.

59 G. Krishnamurthy and R. Turner: “Data Comparison AGC System for VSB Receiver”, U. S. patent No. 5 841 820, 24 November 1998.

signals, which can be lowpass filtered to develop AGC signal. In a variant of this AGC technique, the further ROM storing LUTs for square roots is dispensed with and the sum of the squares is lowpass filtered to develop AGC signal.

6.3.8 Carrier-Derived AGC

The zero-frequency response to the pilot carrier can be extracted from the demodulated baseband DTV signal and used as a basis for developing AGC.[60] Since the amplitude of the pilot carrier is unaffected by suppressed-carrier VSB modulation, this form of AGC can hold gain constant despite modulation not being random. The problem with this form of AGC is that multipath distortion is apt to affect differently the amplitudes of pilot carrier and portions of the signal distal from carrier. Selective cancellation of pilot carrier by multipath generates an AGC signal that increases receiver gain too much, causing non-linear response to the portions of the signal distal from carrier.

Accordingly, if this form of AGC is used, it generally supplements some other type of AGC that forestalls receiver gain being increased enough to cause non-linear response to the portions of the signal distal from carrier. Carrier derived AGC signal can be used to mitigate the effects of different forms of modulation on AGC signal developed from the average amplitude of data, for example.

6.3.9 AGC Developed from Analog-to-Digital-Converter Overflow

In some DTV receivers, which were sold in significant number, analog AGC is based on the detection of overflow from the A/D converter, with gain in the converter and IF amplifier stages of the DTV receiver being reduced sufficiently to avoid the occurrence of such overflow. A problem apt to occur in such an AGC arrangement is that recurrent impulse noise is apt to reduce gain in the receiver more than necessary or desirable. There is reduction of gain for symbols that are not so corrupted by the impulse noise as to be undetectable, so that these symbols are digitized with fewer bits resolution. Presuming there is no inversion of strong impulse noise before the A/D converter, permitting A/D converter overflow on strong impulse noise is desirable. This is because the concurrent symbols are so corrupted as to be undetectable anyway, and symbols that are detectable have the benefit of spanning a greater portion of the A/D converter dynamic range.

Inversion of strong impulse noise in circuitry before the A/D converter can help avoid the problem of AGC based on detection of overflow from the converter inappropriately reducing gain in the converter and IF amplifier stages of the DTV receiver. Arranging the AGC to have slow-attack, fast-release characteristics can also be helpful with these problems.

6.3.10 AGC from Distribution of Analog-to-Digital-Converter Response

AGC can be predicated on a histogram of recent A/D converter response. This has been done indirectly, by keeping track of data slicing results. If the received signal is a trellis-coded 8-VSB signal, except for the initial data segments of each field, all eight of the possible data slicing results should occur nearly the same number of times on average. If the data slicing results for low-amplitude modulation occur a disproportionately large portion of the time, receiver gain is increased. If the data slicing results for high-amplitude modulation occur a disproportionately large portion of the time, receiver gain is decreased.

[60] C. B. Patel and A. L. R. Limberg: "Automatic Gain Control of Radio Receiver for Receiving Digital High-Definition Television Signals", U.S. patent No. 5 636 252, 3 June 1997.

AGC based on a histogram of data slicing results is best employed as a "touch-up" AGC because severe receiver overload may make the data slicing results an inaccurate way of measuring receiver gain. This type of AGC is sometimes used as a "touch-up" AGC controlling the gain of a two-quadrant analog multiplier preceding the A/D converter, so the converter operates over correct dynamic range. Alternatively, this type of AGC is sometimes used as a "touch-up" AGC entirely within the digital domain, using a digital multiplier after the A/D converter as a gain control element. This arrangement does not protect the A/D converter from being operated outside correct dynamic range, however.

AGC based on a histogram of data slicing results does not work for PAM with fewer than four modulation levels.

6.3.11 AGC from Envelope Detection of IF Signal

The IF signal may be envelope-detected, and an average detector may then be used to determine the average value of the detected envelope to develop the basis for AGC. The peak value of modulation is inferred from the average value of the detected envelope, based on knowledge of the signal statistics. This makes the AGC less apt to be affected by impulse noise than if a peak detector is used to determine the peak value of the detected envelope. The envelope of the signal boosted by the presence of pilot carrier is usually the one that is detected, so that the average-value detector averages out Johnson noise.

6.3.12 Differences in Delayed AGC for ATSC and NTSC Receivers

For NTSC reception, the RF AGC delay point is typically set at 0 dBmv in order to achieve a good—perhaps 50 dB—picture-to-noise ratio for analog demodulation. However, DTV does not require as high a signal-to-noise ratio, so the delay point may be set differently. The delay point may be set at a lower received signal level to affect a better compromise that yields RF and mixer linearity for reducing cross- or inter-modulation distortion from strong interfering stations.

6.4 Analog-to-Digital Conversion

In DTV receivers in which the DTV signal is demodulated in analog circuitry, analog-to-digital conversion is performed in the analog regime on the baseband signal, following the example of the Grand Alliance receiver [3]. In most later-developed designs analog-to-digital conversion is performed on an IF signal and the digitized IF signal is demodulated in the digital regime.

Good performance of a DTV receiver under multipath reception conditions where the principal signal has little more energy than its stronger echoes depends upon the digitized signal exhibiting relatively little quantization error. The A/D converter is usually designed to exhibit more than eight bits of resolution in its output signal. When demodulation is done in the digital regime, rather than being done in the analog regime, more bits of resolution in the A/D converter output signal are preferred for reducing quantization error associated with the digital demodulation.

In order to minimize ISI, it is necessary for the A/D converter to sample the DTV signal at twice the Nyquist rate, at least, if root-raised-cosine digital filtering is to be performed subsequently. If root-raised-cosine filtering is performed prior to digitization of the DTV signal, the A/D converter can sample as low as the Nyquist rate without ISI being generated in the analog-to-digital conversion process.

6.4.1 Polyphase Analog-to-Digital Conversion

Polyphase analog-to-digital conversion permits successive-approximation A/D converters of the sigma-delta type to be used instead of flash (ladder) converters.[61] Sigma-delta A/D converters can provide 12-16 bits of digital resolution more easily than a flash converter can. Polyphase analog-to-digital conversion of baseband DTV signal supports polyphase digital filtering to obtain the root-raised-cosine filter response necessary to suppress intersymbol interference (ISI). Polyphase analog-to-digital conversion of IF DTV signal is apt to be followed by demodulation performed on a polyphase basis before root-raised-cosine filtering performed on a polyphase basis. Polyphase root-raised-cosine filtering is usually followed by decimation of baseband DTV signal to twice symbol rate (or in some instances symbol rate) if baseband equalization filtering is used.

6.4.2 Analog-to-Digital Conversion of IF Signals

Most DTV receiver designs developed between 1996 and 2002 digitize IF DTV signals, rather than baseband DTV signals. One reason for this is that it is easier to avoid incurring non-linearity when controlling the gain of input signals to the analog-to-digital converter.

Some digital communications receivers digitize at IF to facilitate equalization being done prior to demodulation being done by synchrodyning, the advantage being that the synchrodyning is less affected by multipath. The disadvantage of equalization before demodulation is that it requires a complex-signal equalizer. A complex-signal equalizer tends to be more expensive in terms of digital multiplications to be performed than a baseband equalizer operative only on real signal, so it is disfavored for designs of DTV receivers for the consumer market.

6.4.2.1 Digitization of Lowband VSB IF a few MHz above Zero Frequency

Digitized IF signals are usually demodulated using complex digital multiplication, with the digitized IF signal being converted to complex form using a digital phase-splitter and then being multiplied by a complex carrier in a synchrodyning procedure. A common type of phase-splitter Hilbert-filters its real-only input signal to generate an imaginary output signal and delays its real-only input signal to supply a real output signal in temporal alignment with its imaginary output signal. The design of the phase-splitter is simplified if the signals to be transformed are in passband form at lowband intermediate frequencies at least 1 MHz above zero-frequency. This avoids having to deal with discontinuity in amplitude or phase of phase-splitter response at zero frequency.

If the lowband VSB IF signal is arranged so its carrier frequency is lower in frequency than the full sideband of that signal, the carrier frequency is usually chosen to be a submultiple of the symbol rate.[62] If the carrier frequency of the lowband VSB IF signal is higher in frequency than the full sideband of the lowband VSB IF signal, the carrier frequency is usually chosen to be a vulgar fraction of the symbol rate, greater than one-half of the symbol rate. These choices of carrier frequency facilitate the generation of digital samples of orthogonal cissoidal carrier signals from look-up tables stored in ROM. The ROM is addressed from sample counters that count samples supplied from a system clock counting samples generated at a rate that is proportional to symbol rate. The digital carrier signals are used in the demodulation or de-rotator procedures carried out in the digital domain.

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[61] A. L. R. Limberg: "Digital TV Receivers with Poly-Phase Analog-to-digital Conversion of Baseband Symbol Coding", U. S. patent No. 5 852 477, 22 December 1998.

[62] C. B. Patel and A. L. R. Limberg: "Digital VSB Detector With Final I-F Carrier At Sub-Multiple Of Symbol Rate, as for HDTV Receiver", U. S. patent No. 5 606 579, 25 February 1997.

Different types of tuner front-end can be used to supply lowband VSB IF signal for digitization. A triple-conversion tuner up-converting received VSB RF signal to a VSB UHF first IF signal, then down-converting amplified VSB UHF first IF signal to a VSB VHF second IF signal, and finally down-converting the VSB VHF second IF signal to a VSB lowband third IF signal has been used.[63] A dual-conversion tuner up-converting received VSB RF signal to a VSB UHF first IF signal and then down-converting amplified VSB UHF first IF signal directly to a VSB lowband second IF signal is also possible.[64]

6.4.2.2 Digitization of VSB DTV IF Signal in VHF Range

A VHF IF signal can be digitized by an A/D converter sampling at a multiple of baud rate. The sampling is done at a rate that is at least symbol rate and is usually a multiple of symbol rate. The sampling time should not be more than a quarter or so of the VHF cycle. VHF IF signal frequencies are chosen so as to heterodyne with the A/D converter rate without introducing aliasing into portions of the spectrum that are to be further processed. The digitized signal is phase-split and demodulated within the digital regime. This can be done by digital synchronous detection that uses carriers supplied from a digital controlled oscillator (DCO), as described in Section 6.7.3 ("Synchronous Detection").

6.4.2.3 ZIF Designs Digitizing VSB IF with Carrier at a Multiple of Baud Rate

If the carrier frequency of an analog VSB IF signal is chosen to be a multiple of the 10.76 MHz baud rate, analog-to-digital conversion of that signal can generate a complex baseband DTV signal directly, without needing complicated filters for digital phase-splitting. This complex baseband DTV signal is sometimes referred to as a "zero-frequency intermediate-frequency" or "ZIF" signal.

If the IF band is much below the conventional 42–48 MHz, there is increased possibility of an image rejection problem when UHF-channel DTV signals are being received. In a DTV receiver with dual conversion to a VHF IF signal, the initial up-conversion to a UHF IF signal avoids image rejection problems when receiving UHF-channel DTV signals. So, the second-IF signals can be lower in frequency.

Choosing too low a carrier frequency poses problems with the A/D conversion harmonics aliasing into baseband when conversion to baseband is done directly as part of the A/D conversion process. A double-conversion tuner can place the full sideband above carrier frequency to mitigate these problems. Second-IF signals with carrier frequencies of 20.52 MHz can be synchrodyned to baseband by A/D converters sampling at 20.52 Msamples/second with their respective sampling phases staggered in time. The two A/D converters respectively generate the real and imaginary components of a complex baseband DTV signal.

Second-IF signals with carrier frequencies as low as (2/3) 10.52 MHz, or so, can be used if the full sideband is above carrier frequency, as can be arranged using a plural-conversion tuner. Second-IF signals with carrier frequencies as low as (3/2) 10.52 MHz, or so, can be used if the full sideband is below carrier frequency. In either case, though, the heterodyne to ZIF or synchrodyne to baseband has to be done by a de-rotator using complex multiplication, to avoid aliasing problems that occur when these procedures are combined with A/D conversion. Digitization of a final IF with a 43.04 MHz carrier provides for a special case of digital demodulation, as outlined in Section 6.7.7 ("Digital Demodulation of IF Signal with Carrier at

[63] C. B. Patel and A. L. R. Limberg: "Digital VSB Detector with Bandpass Phase Tracker, as for Inclusion in an HDTV Receiver", U. S. patent No. 5 479 449, 26 December 1995.

[64] A. L. R. Limberg and C. B. Patel: "Digital TV Signal Receiver with Direct Conversion from UHF I-F to Low-Band I-F Before Digital Demodulation", U. S. patent No. 6 496 230, 17 December 2002.

Even Harmonic of Symbol Rate"), *infra.* Only a single A/D converter is needed for implementing this special case of digital demodulation.

6.4.2.4 Synchronous Digitization of VSB IF in 37.35 to 43.35 MHz Range

A DTV receiver can use either single conversion or double conversion for translating received VSB DTV signal to a VHF IF band with a 43.04 MHz carrier frequency, which is eight times 5.38 MHz, and with its full sideband below carrier frequency. Digital carriers sampling orthogonal phases of the 43.04 MHz carrier frequency can be generated from look-up tables stored in ROM and addressed by a counter clock counting at a multiple of baud rate. Such digital carriers can then be used signal for synchrodyning phase-split VHF DTV IF signal to real and imaginary baseband DTV signals. Alternatively, the digital carriers descriptive of 43.04 MHz carrier frequencies can be used for down-converting VHF DTV IF signal to a complex zero-intermediate frequency signal. This ZIF signal is then demodulated by vector addition of its real and imaginary components.

6.4.2.5 Synchronous Digitization of VSB IF in 42.73 to 48.73 MHz Range

A DTV receiver can use double conversion for translating received VSB DTV signal to a VHF IF band with a 43.04 MHz carrier frequency, which is eight times 5.38 MHz, and with its full sideband above carrier frequency. A DTV receiver can use either single conversion or double conversion for translating received VSB DTV signal to a VHF IF band with a 48.42 MHz carrier frequency, which is nine times 5.38 MHz, and with its full sideband below carrier frequency. Digital carriers sampling orthogonal phases of the chosen carrier frequency can be generated from look-up tables stored in ROM and addressed by a counter clock counting at a multiple of baud rate. Such digital carriers can then be used for synchrodyning phase-split VHF DTV IF signal to real and imaginary baseband DTV signals.

Alternatively, such digital carriers can be used for down-converting VHF DTV IF signal to a complex zero-intermediate frequency signal. This ZIF signal is then demodulated by vector addition of its real and imaginary components.

6.5 Channel Filtering

Root-raised-cosine filtering is required in the receiver, if intersymbol interference (ISI) is to be minimized. Filtering should be done to minimize ISI in the baseband DTV signal supplied to adaptive equalization filtering. Otherwise, ISI can interfere with optimal performance of the auto-regression techniques used for adaptation of the equalizer weighting coefficients.

Additional carrier-side root-raised-cosine roll-off of the IF signal offered for demodulation is desirable, so carrier-side raised-cosine roll-off of the IF signal is established. If the received signal has little or no multipath distortion, carrier-side raised-cosine roll-off of the IF signal causes the real component of demodulated DTV signal to have flat amplitude response extending down from middle frequencies to zero frequency. Flat amplitude response extending down to zero frequency prevents droop in the reproduction of low-frequency transients associated with occasional non-randomness of the transmitted trellis-coded data. Such droop can cause error in equalizer adaptation. Carrier-side raised-cosine roll-off of the IF signal tends to reduce the burden on the adaptive equalizer to provide equalization of low frequencies and tends usually to speed up initial convergence of adaptive equalizer weighting coefficients to correct values.

The raised-cosine roll-off of the VSB DTV signal required to establish Nyquist slope can be developed by additional root-raised-cosine roll-off of the IF signal at the channel edge remote from the carrier. Alternatively, the additional root-raised-cosine roll-off of the DTV signal to develop the raised-cosine roll-off required to establish Nyquist slope can introduced by filtering of the baseband DTV signal resulting from demodulation.

6.5.1 Root-Raised-Cosine IF SAW Filtering to Establish Nyquist Slope

Root-raised-cosine filtering of the side of the VSB IF spectrum remote from carrier can be done by SAW filtering in the IF amplifier circuitry. However, there is some difficulty assuring that the VSB IF spectrum is aligned properly within the SAW filtering. This frequency alignment problem is avoided by performing the root-raised-cosine filtering to establish Nyquist slope on the baseband DTV signal, rather than establishing Nyquist slope by filtering the IF DTV signal.

6.5.2 IF SAW Filtering Establishing Carrier-Side Raised-Cosine Roll-Off

Root-raised-cosine filtering of the side of the VSB IF spectrum close to carrier can be done by SAW filtering in the IF amplifier circuitry. However, such filtering reduces pilot carrier energy, reducing capability for stabilizing phase lock of the AFPC loop used to control local oscillator frequency in synchrodyning procedures used for demodulating the DTV signal.

6.5.3 Digital Nyquist Filtering of IF Signal

Digital filtering of IF signal for establishing root-raised-cosine filtering of both sides of the VSB IF spectrum is known.[65]

Lowpass digital filtering of a digitized lowband-IF signal with principal sideband located below a carrier frequency that is a prescribed fraction of baud rate can be done to establish carrier-side roll-off of channel frequency response. This is followed by digital demodulation and then by lowpass digital filtering of the baseband DTV signal for establishing Nyquist slope. This method reduces the number of samples over which the digital filtering need be done to optimize response of the baseband DTV signal near zero frequency. Use of this method with a bandpass tracker can avoid the need for automatic fine tuning in preceding portions of the DTV receiver.

Highpass digital filtering of a digitized lowband-IF signal with principal sideband located above a carrier frequency that is a prescribed fraction of baud rate can be done to establish carrier-side roll-off of channel frequency response. Digital demodulation and then lowpass digital filtering of the baseband DTV signal for establishing Nyquist slope follow the highpass digital filtering. This method reduces still further the number of samples over which the digital filtering need be done to optimize response of the baseband DTV signal near zero frequency. Sampling rate need only be high enough to support lowpass digital filtering of the baseband DTV signal for establishing Nyquist slope, a sampling rate four times baud rate being sufficient.

Use of this method with a bandpass tracker can avoid the need for automatic fine tuning (AFT) in preceding portions of the DTV receiver. Digital filtering of the digitized lowband-IF signal can be disabled if tracking has not been established by the automatic-frequency-and-phase control (AFPC) loop for the local oscillator in the bandpass tracker. This reduces the possibility of AFPC lock-out.

6.5.4 Root-Raised-Cosine Baseband Filtering to Establish Nyquist Slope

The root-raised-cosine filtering to establish Nyquist slope can be done by digital lowpass filtering of baseband DTV signal, as supplied directly from digital demodulation circuitry, or as supplied from an A/D converter digitizing the analog baseband DTV signal supplied from analog demodulation circuitry. The digitized baseband DTV signal must be sampled at twice baud rate

[65] W. Boie and E. B. Sieman: "Circuit for Carrying Out Digital Nyquist Filtering of IF Intermediate Frequency Signals", U. S. patent No. 5 963 273, 5 October 1999.

or at higher rate. Practical designs involve polyphase digital filtering with clocking frequencies of four or eight times symbol rate. This is a self-aligning way of establishing the Nyquist slope that is unaffected by a DTV signal carrier frequency being offset from nominal value.

6.5.5 Baseband Filtering to Establish Desired Response Near Zero-Frequency

If the side of the VSB IF spectrum close to carrier is filtered to exhibit a raised-cosine roll-off 6 dB down at carrier frequency, the baseband DTV signal will have a flat amplitude response down to zero-frequency. If little or no filtering is done to roll off the side of the VSB IF spectrum close to carrier, folding of the spectrum around zero-frequency during synchronous demodulation introduces a boost into the low-frequency end of the frequency spectrum of the real baseband DTV signal. Synchronous demodulation also generates a zero-frequency component of the real baseband DTV signal, which zero-frequency component results from pilot carrier detection. Digital highpass filtering of the real baseband DTV removes the zero-frequency component and flattens the low-frequency boost.

6.6 Symbol Clock Recovery and Symbol Synchronization

6.6.1 Bright-Spectral-Line Techniques

Because of the VSB DTV signal being accompanied by pilot carrier, envelope detection of the received DTV signal provides a signal having a strong frequency component at one-half baud rate. The envelope detector response is then non-linearly processed to generate frequency components at baud rate and harmonics thereof, one of which components is separated by a narrowband filter. The frequency component separated by the narrowband filter is used for frequency-locking a master clock oscillator with a natural frequency close to the separated frequency component. The master clock oscillator is frequency-locked by *automatic-phase-and-frequency-control* (AFPC) loop in some designs and by injection lock in other designs. A properly designed AFPC loop can better control phase jitter during noisy reception. In still other designs the AFPC loop includes a frequency divider, so the master clock oscillator is frequency-locked to a multiple of the frequency of the narrowband filter response.

While very precise frequency lock of the master clock oscillator is simple to achieve using these bright-spectral-line techniques, accurate phase lock can be problematic. Any variation of the phase response of the narrowband filter in the frequency range immediately surrounding the separated frequency component affects the phase of the master clock. Accordingly, some means of introducing an automatic phase adjustment of the clocking is used together with a bright-spectral-line technique, so that symbols are detected at optimal timing for avoiding ISI. This phase adjustment is essentially a static adjustment, which facilitates its being made in any of numerous ways. One way is to use a phase discriminator on DSS signals, as described in [3, Section 9.2.4].

6.6.2 Qureshi-Type Symbol Synchronization Techniques

After filtering to minimize ISI and to remove zero-frequency component, the baseband DTV signal is sampled at the second harmonic of baud rate in one Qureshi-type symbol synchronization technique. If alternate ones of the samples occur at the center of symbols, intervening samples of the signal theoretically should be zero-valued, because supposedly there is no ISI. The phase of alternate samples of the baseband DTV signal having smaller average amplitude is determined and the departures from zero-value of these samples are noted. These departures are then processed for determining the error in sampling phase, which processing tends to be complicated.

In an alternative Qureshi-type symbol synchronization technique that is simpler to implement, the baseband DTV signal is sampled at a harmonic of baud rate, so there are H phases of samples in each symbol epoch. H can be as low as three, but is usually chosen to be an even number at least four. The one of the H phases of samples having the highest average amplitude is determined. If the average amplitude of the one of the H phases of samples preceding the one with highest energy exceeds the amplitude of the one of the H phases of samples succeeding the one with highest energy, an automatic phase control (APC) signal is generated that retards the sampling clock. If the average amplitude of the one of the H phases of samples succeeding the one with highest energy exceeds the amplitude of the one of the H phases of samples preceding the one with highest energy, an APC signal is generated that advances the sampling clock.

Error in sampling phase is continuously tracked by Qureshi-type symbol synchronization techniques, which can be advantageous for implementing schemes that compensate for symbol jitter during multipath reception.[66]

6.6.3 Band-Edge Filtering Technique

A DTV receiver supplies final-IF VSB DTV signal the carrier frequency of which is nominally at the 5.38 MHz Nyquist frequency to an A/D converter. The A/D converter samples the final-IF VSB DTV signal at a nominal 21.52 Msamples/s. The digitized final-IF VSB DTV signal is supplied to band-edge digital filters 2.69 MHz above and below the nominally 5.38 MHz carrier frequency. These band-edge digital filters are complex filters with cut-offs complementary to the final-IF VSB DTV signal producing a DSB AM signal response in the absence of pilot signal. The response of the carrier-side band-edge filter is convolved with the conjugate of the response of the other band edge filter. The imaginary component of the convolution result is a measure of the mis-timing of the sampling by the A/D converter, and the real component of the convolution result indicates the direction of the mis-timing. Corrections are made using phase detection techniques similar to those used in Qureshi-type symbol synchronization.[67]

6.6.4 PN-Sequence Phase-Discrimination Techniques

A match filter for the PN511 sequence in the DFS signal, which match filter processes baseband DTV signal at a multiple of baud rate, can be used as a symbol phase discriminator. Automatic adjustment of sampling phase is done to place the peak response of the match filter at a sample epoch that is flanked by sample epochs with equal-valued shoulder responses. The sampling phase at which the match filter peaks is then used for reckoning the centers of symbol epochs.

Match filters for the PN63 sequence in the DFS signal can be similarly used, but are more sensitive to noise and echoes of adjoining signal.

6.6.5 Data-Sequence Phase-Discrimination Technique

The symbol clock can be adjusted in response to error in the sampling time of the data segment sync sequence.[68 69 70]

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[66] S. U. H. Qureshi: "Timing Recovery for Equalized Partial-Response Systems", *IEEE Trans. On Communications*, pp. 1326–1330, December 1976.

[67] C. H. Strolle and S. T. Jaffe: "Carrier Independent Timing Recovery System for a Vestigial Sideband Modulated Signal", U. S. patent No. 5 805 242, 8 September 1998.

[68] Key H. Kim: "Data segment sync signal detector for HDTV", U. S. patent No. 5 548 339, August 1996.

[69] Ki-Bum Kim, et al.: "A Symbol Timing Recovery Using The Segment Sync Data for the Digital HDTV GA VSB System," *IEEE Trans. on Consumer Electronics*, vol. 42, no. 3, pp. 651–655, August 1996.

[70] K. B. Kim: "Symbol Timing Recovery Circuit and Method", U. S. patent No. 5 859 671, 12 January 1999.

6.6.6 Techniques Adjusting Twice-Baud-Rate Clock so Odd and Even Samples of Baseband DTV Signal Pair up on Average

The symbol clock can be adjusted so that odd and even twice-baud-rate samples pair up on average.[71] [72]

6.6.7 Lowest-Slicer-Error Technique

An adaptive equalizer that supplies response at a multiple of baud rate is asynchronously clocked. Its response is decimated in a number (e.g., 16) different ways with each of the decimated responses being a baud-rate signal data sliced by a respective data slicer. A histogram of the cumulative decision error from each data slicer is generated, and the decimated equalizer response data slicer generating the least cumulative decision error is selected for trellis decoding.

6.6.8 Auxiliary Adaptive Equalizer for Determining Optimum Symbol Phasing

The optimum positioning of the reference tap of a principal adaptive equalizer can be determined using an auxiliary adaptive equalizer.[73] The auxiliary adaptive equalizer successively tries sampling with different phasing. In each successive try with different phasing, the weighting coefficients of the auxiliary the adaptive equalizer are adapted using an adaptation algorithm. Then, the reception errors experienced by the two adaptive equalizers are accumulated, and the respective accumulation results are compared. If the accumulated reception error experienced by the auxiliary adaptive equalizer is smaller than the accumulated reception error experienced by the principal adaptive equalizer, the weighting coefficients of the principal adaptive equalizer are modified to correspond with the weighting coefficients of the auxiliary adaptive equalizer.

6.7 Demodulation and Preparation of VSB Signal for Demodulation

Most DTV receivers designed between 1996 and 2000 performed demodulation using digital techniques, rather than using analog synchronous detection as done in the Grand Alliance receiver [3]. Sometimes, demodulation is referred to as "de-rotation".

6.7.1 Phase Splitting Prior to Demodulation

6.7.1.1 Digital-Regime Phase-Splitters

Digital phase-splitter filtering is used to convert a digital real-only DTV signal to a complex DTV signal that can be synchronously detected by multiplying it with a complex digital carrier having a cissoidal system function. A commonly used digital phase-splitter uses a Hilbert filter for generating an imaginary DTV signal from the real-only DTV signal and delays the real DTV signal to align it temporally with the imaginary DTV signal.

Another type of digital phase-splitter filtering first described by Rader has been considered for use in a DTV receiver design.[74] [75] Yet another type of digital phase-splitter filtering first described by Ng has been considered for use in a DTV receiver design.[76] [77]

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[71] Y. S. Choi: "Apparatus for Recovering Full Digital Symbol Timing", U. S. patent No. 5 872 818, 16 February 1999.

[72] S. S. Oak, W. J. Song, and K. C. Lee: "Maximum Likelihood Symbol Timing Estimator", U. S. patent No. 6 341 147, 22 January 2002.

[73] H. Sari: "Method of and Arrangement for Determining the Optimum Position of the Reference Tap of an Adaptive Equalizer", U. S. patent No. 4 633 482, 30 December 1986.

[74] C. M. Rader: "A Simple Method for Sampling In-Phase and Quadrature Components", *IEEE Trans. on Aerospace & Electronic Systems*, vol. AES-20, no. 6, pp. 821–824, November 1984.

6.7.1.2 Analog-Regime Phase-Splitters

Analog phase-splitter filtering is known.[78] Because separate A/D converters are required for the real and imaginary components of the analog complex DTV signal, designers usually do not favor phase-splitting in the analog regime. Also, the tracking of the real and imaginary components of the analog complex DTV signal is perhaps harder to assure in the analog regime.

6.7.2 VSB-to-DSB-AM Conversion before Demodulation

Conversion of the VSB DTV signal to a double-sideband amplitude-modulation (DSB-AM) signal prior to demodulation is simplest to do in the digital regime, although analog methods are also possible. The digital VSB-to-DSB AM conversion proceeds from a digital VSB IF signal with carrier frequency related to the symbol rate by prescribed whole-number ratio greater than one-half. The complex digital VSB IF signal is heterodyned with complex beat-frequency oscillations at twice carrier frequency to generate a reversed-spectrum complex digital VSB IF signal. The forward- and reversed-spectrum VSB IF signals are combined to generate in-phase and quadrature-phase DSB-AM signals for demodulation. The VSB-to-DSB AM conversion can cancel phase noise and so reduce symbol jitter.

6.7.3 Synchronous Detection

Demodulation in the analog regime has some advantages over demodulation in the digital regime. If square-root-raised-cosine filtering is performed in the IF amplifier chain, analog demodulation using synchronous detection techniques (e.g., a quadricorrelator) offers precision beyond that obtained with ordinary digital demodulation and uses less hardware. The analog baseband DTV signal can then be digitized at a sampling rate as low as symbol rate and that need be no higher than twice symbol rate.

Synchronous detection is usually done in the digital regime in most designs of DTV receivers for the consumer market, however. One reason for this is that digital synchronous detection is more easily incorporated into an integrated-circuit design. Digital synchronous detection is usually complex in nature, using a digital complex multiplier to multiply complex IF DTV signal from phase-splitter filtering by a complex carrier signal supplied from a *digital controlled oscillator* (DCO).[79] [80] [81] The DCO generally comprises sine/cosine tables stored in ROM addressed by an accumulator for sample count, as incremented or decremented by current measurement of error.

Although the automatic phase control (APC) loop for the DCO used first-order digital loop filtering in early designs, a later design successfully employs higher-order digital loop filtering

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[75] C. B. Patel and A. L. R. Limberg: “Digital VSB Detector With Bandpass Phase Tracker Using Rader Filters, as for Use in an HDTV Receiver”, U. S. patent No. 5 548 617, 20 August 1996.

[76] T. F. S. Ng: “Quadrature Demodulator” U. K. patent application 2 244 410A, 27 November 1991.

[77] C. B. Patel and A. L. R. Limberg: “Digital VSB Detector with Bandpass Phase Tracker Using Ng Filters, as for Use in an HDTV Receiver”, U. S. patent No. 5 731 848, 24 March 1998.

[78] A. L. R. Limberg: “Bandpass Phase Tracker with Hilbert Transformation before Plural-Phase Analog-to-digital Conversion”, U. S. patent No. 5 982 820, 9 November 1999.}

[79] T. P. Horwitz, R. B. Lee, and G. Krishnamurthy: “Error Tracking Loop”, U.S. Patent No. 5 406 587, 11 April 1995.

[80] G. Krishnamurthy and R. B. Lee: “Simplified Complex Multiplier in Error Tracking Loop”, U. S. patent No. 5,533,070, 2 July 1996.

[81] G. Krishnamurthy and R. B. Lee: “Error Tracking Loop Incorporating Simplified Cosine Look-up Table”, U. S. patent No. 5 533 071, 2 July 1996.

that can integrate out static phase error.[82] The integrator of the higher-order digital loop filtering setting up on noise is detected, and the integrator output signal is reset to prescribed fixed value momentarily converting loop filtering to first-order. This speeds the response of the APC loop when noise is present.

6.7.4 Bandpass Phase Tracker

A bandpass phase tracker uses a local oscillator with automatic frequency and phase control (AFPC) for supplying the oscillations heterodyned with VHF (or UHF) IF signals to generate a lowband IF signal for demodulation in the digital regime. Demodulation is done by a complex synchronous detector supplied a complex digital carrier generated from look-up tables stored in ROM. The ROM is addressed by a sample counter counting samples at a fixed rate that is at least as high as symbol rate and is proportional thereto. The AFPC signal is generated from the imaginary output signal such demodulation develops, either by simply lowpass filtering the imaginary output signal[83] or by using a Costas loop. In one species of Costas loop, the real output signal from demodulation is highpass filtered, symmetrically clipped, and used to multiply the imaginary output signal from demodulation to generate a product signal that is lowpass filtered to generate the AFPC signal.[84] In another species of Costas loop, the real and imaginary output signals from demodulation are squared. Lowpass filter response to the product of multiplying the squaring results together provides the AFPC signal.[85]

Demodulation is carried out by a complex digital multiplier multiplying phase-split lowband IF signal by a digitized complex carrier. The digitized complex carrier is customarily generated from sine/cosine tables stored in ROM addressed by an address counter counting transitions in a clocking signal that is a multiple of baud rate. The fact that the digital carrier waves are not altered during the course of demodulation is one reason for using the bandpass tracker. Another reason the bandpass tracker is used is that digital phase-splitting can be done very accurately at lowband IF. If the lowband IF signal is a VSB signal in which the sideband with frequencies below the carrier are mostly suppressed, in representative bandpass tracker designs its carrier is located at one-half, one-quarter, one-eighth and one-sixteenth baud rate, respectively. If the lowband IF signal is a VSB signal in which the sideband with frequencies above the carrier are mostly suppressed, its carrier is located no lower than one-half baud rate. In representative bandpass tracker designs the carrier is located at two-thirds baud rate, at three-quarters baud rate, and at baud rate, respectively.[86] [87] [88]

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82 B. B. Bhatt: "Higher Order Digital Phase Loop Filter", U. S. patent No. 5 774 512, 30 June 1998.

83 C. B. Patel and A. L. R. Limberg, "Radio Receiver for Vestigial-Sideband Amplitude-Modulation Digital Television Signals", U. S. patent No. 6,512,555, 28 January 2003.

84 D. S. Han: "Digital Carrier Wave Restoring Device and Method for Use in a Television Signal Receiver", U. S. patent No. 5 809 088, 15 September 1998.

85 K. Kimura: "Costas Loop Carrier Recovery Circuit Using Square-law Circuits", U. S. patent No. 5 982 200, 9 November 1999.

86 C. B. Patel and A. L. R. Limberg: "Digital VSB Detector with Bandpass Phase Tracker, as for Inclusion in an HDTV Receiver", U. S. patent No. 5 479 449, 26 December 1995.

87 C. B. Patel and A. L. R. Limberg: "Digital TV Detector Responding to Final-IF Signal with Vestigial Sideband Below Full Sideband in Frequency", U. S. patent No. 5 659 372,19 August 1997.

88 A. L. R. Limberg: "Bandpass Phase Tracker with Hilbert Transformation Before Plural-Phase Analog-To-Digital Conversion", U. S. patent No. 5 982 820, 9 November 1999.

6.7.5 Bandpass Phase Tracker with VSB AM to DSB AM Conversion

A DSB AM lowband IF signal can be generated as the down-conversion result of heterodyning a VHF (or UHF) DTV IF signal with the respective oscillations from a pair of local oscillators, rather than heterodyning the DTV IF signal with the oscillations from a single local oscillator. The respective oscillations from the pair of local oscillators should differ in frequency by twice the carrier frequency of the DSB AM lowband IF signal. Oscillations from one of the pair of local oscillators are higher in frequency than the DTV IF signal, so there is spectrum reversal of the DTV signal in its component of the down conversion result. Oscillations from the other of the pair of local oscillators are lower in frequency than the DTV IF signal, so there is no spectrum reversal of the DTV signal in its component of the down conversion result.

Rather than actually using the pair of local oscillators to supply the two beat frequencies used in down-conversion, the two beat frequencies may be provided from a balanced modulator that modulates oscillations from one local oscillator by the carrier frequency of the DSB AM lowband IF signal. It is convenient to automatically control the frequency and phase of this one local oscillator similarly to the way the frequency and phase of the local oscillator in a bandpass tracker is controlled, as described in Section 6.7.4 (“Bandpass Phase Tracker”). The balanced modulator portions of the VSB AM to DSB AM conversion are implemented entirely within the digital regime in designs made in the late 1990s.

6.7.6 Quasi-Synchronous Detection Employing Zero-Intermediate-Frequency (ZIF)

A complex ZIF signal is the down-conversion result of mixing a UHF or VHF DTV signal with a local oscillation signal similar in frequency to the carrier frequency of that DTV signal in a tentative demodulation procedure. The local oscillation signal can be from a crystal-controlled oscillator operating as a beat-frequency oscillator or can be from a local oscillator having its frequency controlled by the received DTV signal. A quasi-synchronous detector does not require phase-splitter filtering of the DTV signal before its conversion to a complex ZIF signal used as a tentative demodulation result. In most quasi-synchronous detection designs the ZIF signal is generated in the digital domain or, if generated in the analog domain, is digitized before further processing.

In some further demodulation procedures for ZIF signal, the real and imaginary components of the ZIF signal are added vectorially to reproduce a baseband DTV signal. In one further demodulation procedure, the real and imaginary components of the ZIF signal are squared and added, with sum of the squares being square-rooted to recover the baseband DTV signal. In alternative further demodulation procedures, various phase-error-detecting (e.g., CORDIC) techniques are used to recover the baseband DTV signal.[89] [90] [91] [92] The further demodulation procedure can also be implemented entirely in ROM addressed by the real and imaginary components of the ZIF signal.

A quasi-synchronous detector is known in which the ZIF error is detected and used for correcting the real component of the ZIF signal. The DTV receiver front-end supplies final-IF VSB DTV signal the carrier frequency of which is nominally at the 5.38 MHz Nyquist frequency

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[89] A. D. Kucar: “Method and Apparatus for Carrier Synchronization and Data Detection”, U. S. patent No. 5 115 454, 19 May 1992.

[90] M. H. Lee: “Phase Detecting Method and Phase Tracking Loop Circuit for a Digital Vestigial Sideband Modulation Communication Device”, U. S. patent No. 5 796 786, 18 August 1998.

[91] M. H. Lee: “Phase Detecting Method and Phase Tracking Loop Circuit for a Digital Vestigial Sideband Modulation Communication Device”, U. S. patent No. 5 933 460, 3 August 1999,

[92] S. Soga, D. Hayashi, T. Hayashi, and S. Sakashita: “Digital Demodulation with Compensation for Phase and Frequency of Tentatively Demodulated Signal”, U. S. patent No. 5 920 228, 6 July 1999.

to an A/D converter. The A/D converter samples the final-IF VSB DTV signal at a nominal 21.52 Msamples/s. The digitized final-IF VSB DTV signal is supplied to band-edge digital filters 2.69 MHz above and below the nominally 5.38 MHz carrier frequency. These band-edge digital filters are complex filters with cut-offs complementary to the final-IF VSB DTV signal producing a DSB AM signal response in the absence of pilot signal. The complex response of the carrier-side band-edge filter is convolved with the complex response of the other band edge filter to generate a DSB AM signal with carrier at twice the frequency of the carrier frequency of the final-IF VSB DTV signal. This DSB AM signal is heterodyned to baseband to develop the second harmonic of the ZIF error, which is frequency-divided to obtain the ZIF error.[93]

6.7.7 Digital Demodulation of IF Signal with Carrier at Even Harmonic of Symbol Rate

The demodulation of a digitized IF signal with a carrier frequency of 43.04 MHz, four times the 10.76 megabaud rate is a special case. The digitized IF signal is sampled at four times baud rate. The set of odd samples and the set of even samples are separated to de-multiplex the real and imaginary components of the signal. Demodulation of the set of odd samples is done by inverting every other one of them to generate a real component of DTV signal at 21.52 Msamples/s rate. Demodulation of the set of even samples is done by inverting every other one of them to generate an imaginary component of baseband DTV signal at 21.52 Msamples/s rate. AFPC signal for the VCO used to generate sampling clock for the A/D converter can be generated by digital Costas technique or by a Qureshi type of technique, by way of example and counterexample.

Other variations of demodulation of a digitized IF signal with a carrier frequency of four times symbol rate are known.[94]

Complex demodulation of a digitized IF signal with a carrier frequency that is a multiple of symbol rate can be done to accommodate analog-to-digital conversion being done by sigma-delta methods.[95] The analog-to-digital conversion, lowpass filtering to reject out-of-band aliasing from the conversion, and subsequent decimation procedures accomplish the demodulation.

6.8 Sample, Symbol, Data-Segment and Data-Field Counting Arrangements for Receiver Control Purposes

In most (if not all) DTV receiver designs, many operations are controlled by decoders responding to the count from a control counter that counts fractions of symbol epochs occurring within each data frame. Decoding operations are simplified by a control counter construction that comprises stages that count fractions of symbol epochs in each symbol epoch, stages that count modulo-832 the symbol epochs in each data segment, stages that count modulo-313 the data segments within each data field, and a stage that counts modulo-2 the data fields within each data frame.

6.9 Data Field Synchronization

A match filter for the PN511 sequence in the data field sync (DFS) signal will generate a pulse that should be occur during the 519th symbol epoch of the initial data segment of each data field. This pulse can be used to reset the portions of a control counter that count fractions of symbol

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[93] C. H. Strolle and S. T. Jaffe: “Carrier Recovery System for a Vestigial Sideband Modulated Signal”, U. S. patent No. 5 805 242, 13 April 1999.

[94] R. J. Inkol: “Efficient Digital Quadrature Demodulator”, U. S. patent No. 5 504 455, 2 April 1996, provides an excellent review of this art as of 1995 in the “Background of the Invention.”

[95] J. G. Mittel: “Apparatus for Deriving In-phase and Quadrature-phase Baseband Signals from a Communication Signal”, U. S. patent No. 5 787 125, 28 July 1998.

epochs in each data segment and data segments within each data field. The response of the match filter for the PN511 sequence for the halves of symbol epochs exhibiting peak responses can be compared in order to generate a phase control signal that controls the phase of clocking at a multiple of baud rate.

The match filter for the PN511 sequence can be simplified for just correlating the "sign" of the DTV baseband signal with the "sign" of a PN511 sequence.[96]

A match filter for the inverted PN63 sequence in each alternate data field sync (DFS) signal receives input signal as gated by decoder circuitry responsive to the count of fractions of symbol epochs within each data field. The pulse output response from the match filter for the inverted PN63 sequence resets the data frame count in the control counter. The match filter for the inverted PN63 sequence can be preceded by a field comb filter for the initial data segment of each field, which field comb filter suppresses symbols preceding and succeeding the inverted PN63 sequence. This allows the match filter response to be more accurately predicted. This facilitates threshold detection of the match filter response for resetting the data frame count.

6.10 Data Segment Synchronization

6.10.1 Data Segment Synchronization with Impulse-Noise Immunity

Data segment synchronization (DSS) sequences can be line comb filtered from the baseband DTV signal recovered by demodulation. Data segment synchronization (DSS) can be inferred from symbol count and used to develop windowing that reduces the delay structure in the line comb filtering. A correlation filter for the DSS sequence is cascaded with the line comb filter, and the response of the cascaded filtering is threshold detected to develop indications of the beginnings of data segments. These indications are provided even when noise obliterates the most recently received DSS sequences.[97]

6.11 Filtering to Suppress Co-Channel NTSC Interference

When the DTV signal is accompanied by an interfering co-channel NTSC signal, the baseband DTV signal is corrupted by artifacts of the frequency components of that NTSC signal generated by their heterodyning with the DTV signal carrier during the downconversion to baseband. Accordingly, so long as NTSC TV broadcasting continues to use the same frequency channels as ATSC DTV broadcasting, filtering to suppress an interfering co-channel NTSC signal is an important component of a DTV receiver. The IF filtering in the DTV receiver can be designed for selecting against the NTSC audio carrier and most of its FM sideband frequencies. However, the NTSC video carrier, its VSB luma modulation and the chroma modulation of its color subcarrier overlap the DTV signal frequencies. So, suppressing artifacts of these components in the baseband DTV signal presents a more challenging filtering problem.

6.11.1 IF SAW Filtering Rejecting FM Audio Carrier of Co-Channel NTSC Interference

SAW filtering in the IF amplifier circuitry can be designed to reject the FM audio carrier of an interfering co-channel NTSC signal. This permits the use of co-channel NTSC interference filters that reject frequencies near the frequencies of NTSC video carrier and chroma subcarrier, but do not reject frequencies near the NTSC audio carrier. The rejection of the FM audio carrier

[96] K. B. Kim: "Apparatus and Method for Detecting Field Sync Signals and Generating Useable Field Sync Signals in a High Definition Television Receiver", U. S. patent No. 5 877 816, 2 March 1999.

[97] Jian Yang: "Line Sync Detector for Digital Television Receiver", U. S. patent No. 5 594 506, 14 January 1997.

by SAW filtering in the IF amplifier circuitry affects the raised-cosine roll-off of the reception channel, so it introduces some intersymbol interference (ISI).[98]

6.11.2 Partial-Response Filtering of Baseband DTV Signal to Suppress Demodulation Artifacts of Co-Channel NTSC Interference

Comb filtering of the baseband DTV signal by combining it with itself as delayed by an integral number D of symbol epochs can suppress artifacts of the frequency components of an interfering co-channel NTSC signal. The reduced artifacts of the frequency components of the interfering co-channel NTSC signal are discriminated against in a subsequent data-slicing process, owing to quantization effect.

In contrast with the procedure described in [3, Section 9.2.7], wherein comb filtering in which D = 12 is used to provide partial-response post-coding complementary to partial-response pre-coding done at the transmitter, the comb filtering can be used to provide partial-response pre-coding of the DTV signal. Then, after the DTV signal is quantized during data slicing, the symbols are subjected to complementary partial-response post-coding to recover the original symbol information without loss. This complementary partial-response post-coding is provided by ISI-suppression filtering that combines the comb-filtered baseband DTV signal supplied to that filtering with the post-coding response as delayed by the integral number D of symbol epochs, for generating the post-coding response. This combining is a modular combining that reduces the number of levels of modulation in the post-coding response of the ISI-suppression filtering to the same number as in the original baseband DTV signal supplied for comb filtering.[99] [100] [101] The ISI-suppression filtering can be viewed as restoring spectral flatness, so that remnant noise in spectral nulls does not exert undue influence on decision feedback and adaptation of the equalization filtering.

The comb filtering used to suppress artifacts of the frequency components of an interfering co-channel NTSC signal is continuously operative on the baseband DTV signal, including its DFS and DSS components. The ISI-suppression filtering used for partial-response post-coding is recursive. So, to avoid running error in the post-coding response, DFS and DSS symbols are inserted into the response at times when such symbols should occur in the response. The respective performances of the various types of comb filtering that can be used to suppress demodulation artifacts of co-channel NTSC interference all depend in some degree upon the current nature of the interfering NTSC signal. So, a DTV receiver may be designed to select different ones of a plurality of different types of comb filtering for suppressing demodulation artifacts of co-channel NTSC interference at different times.[102] [103] [104]

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98 A. L. R. Limberg: “DTV Receiver with Filter in I-F Circuitry to Suppress FM Sound Carrier of NTSC Co-Channel Interfering Signal”, U. S. patent No. 6 400 393, 4 June 2002.

99 R. W. Citta: “Co-Channel Interference Filter For Digital High Definition Television Receiver”, U. S. patent No. 5 132 797, 21 July 1992.

100 R. W. Citta: “Co-Channel Interference Filter For Television Receiver”, U. S. patent No. 5 162 900, 10 November 1992.

101 A. L. R. Limberg: “DTV Receiver Symbol Decoding Circuitry with Co-Channel NTSC Artifacts Suppression Filter Before Data Slicer”, U. S. patent No. 6 380 969, 30 April 2002.

102 A. L. R. Limberg: “Digital TV Receiver with Adaptive Filter Circuitry for Suppressing NTSC Co-Channel Interference”, U. S. patent No. 5 748 226, 5 May 1998.

103 A. L. R. Limberg: “Digital TV Receiver Circuitry for Detecting and Suppressing NTSC Co-Channel Interference”, U. S. patent No. 5 801,759, 1 September 1998.

104 A. L. R. Limberg: “Digital TV Receiver with Adaptive Filter Circuitry for Suppressing NTSC Co-Channel Interference”, U. S. patent No. 5 835 131, 10 November 1998.

Variable comb filtering has been considered by designers in the attempt to optimize the trade-off between best rejection of co-channel NTSC interference and best AWGN performance.[105] [106] [107]

6.11.2.1 Partial-Response Filtering Using a Comb Filter that Subtracts Baseband DTV Signal as Delayed by Twelve Symbol Epochs from its Undelayed Self

A comb filter that subtracts the baseband DTV signal as delayed by twelve symbol epochs from its undelayed self will suppress artifacts of frequency components of an interfering co-channel NTSC signal near its video carrier, its chroma subcarrier and its audio carrier. If this form of comb filter is used for suppressing demodulation artifacts from an interfering co-channel NTSC signal, twelve-phase trellis decoding of the comb-filtered baseband DTV signal can be done with complete independence between phases.[108]

In this form of comb filter as originally described, the subtraction of the baseband DTV signal as delayed by twelve symbol epochs from its undelayed self was performed on a modular basis. Modular subtraction is infeasible when there is substantial multipath distortion of the DTV signal. So, more recent embodiments of this form of comb filtering use simple subtraction. Modularization of the comb-filtered DT signal is deferred until after equalization, generally being performed in the data-slicing procedures associated with trellis coding and with decision-feedback.

6.11.2.2 Partial-Response Filtering using a Comb Filter that adds Baseband DTV Signal to Itself as Delayed by Six Symbol Epochs

A comb filter that adds the baseband DTV signal to itself as delayed by six symbol epochs will suppress artifacts of frequency components of an interfering co-channel NTSC signal near its video carrier and its chroma subcarrier. As compared to a comb filter that differentially combines the DTV signal delayed by twelve symbol epochs with its undelayed self, the dips in the filter response near NTSC video carrier and chroma subcarrier frequencies are doubled in width. So, it is probable that more of the high-energy content of the interfering co-channel NTSC signal will be suppressed by quantization effect during data-slicing. Furthermore, the correlation between NTSC video artifacts differentially delayed by six symbol epochs is likely to be better than between NTSC video artifacts differentially delayed by twelve symbol epochs.

6.11.2.3 Partial-Response Filtering of DTV Signal using NTSC-Line-Comb Filtering

Partial-response filtering of DTV signal using NTSC-line-comb filtering has been considered. A line-comb filter that subtracts the baseband DTV signal as delayed by 684 symbol epochs from its undelayed self will suppress artifacts of the video carrier and luminance components of an interfering co-channel NTSC signal. A line-comb filter that adds the baseband DTV signal as delayed by 684 symbol epochs to its undelayed self will suppress artifacts of the color burst and chrominance components of an interfering co-channel NTSC signal. A two-line comb filter that subtracts the baseband DTV signal as delayed by 1368 symbol epochs from its undelayed self is

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105 R. W. Citta: "Co-Channel Interference Filter for Digital High Definition Television Receiver", U. S. patent No. 5 132 797, 21 July 1992.

106 R. W. Citta: "Co-Channel Interference Filter For Television Receiver", U. S. patent No. 5 162 900, 10 November 1992.

107 S. N. Hyulyalkar: "Method And Apparatus for Combating Co-Channel NTSC Interference Using a Variable-Comb Filter for Digital TV Transmission", U. S. patent No. 5 648 822, 15 July 1997.

108 R. W. Citta, D. M. Mutzabaugh, and G. J. Sgrignoli: "VSB HDTV Transmission System with Reduced NTSC Co-Channel Interference", U. S. patent No. 5 087 975, 11 February 1992.

equivalent to a cascade connection of these two line-comb filters. This two-line comb filter is suitable for partial-response filtering of DTV signal.

The principal problem with partial-response filtering using NTSC-line-comb filtering is that a "precipitous" horizontal edge in the NTSC video image being transmitted can cause demodulation artifacts of an interfering co-channel NTSC signal to be poorly suppressed for a few hundred symbol epochs. This imposes a burden on the Reed-Solomon error-correction circuitry that is usually significantly greater than the burden imposed by partial-response filtering that combines baseband DTV signal differentially delayed by a small number of symbol epochs, as described in Section 6.11.2.1 ("Partial-Response Filtering Using a Comb Filter that Subtracts Baseband DTV Signal as Delayed by Twelve Symbol Epochs from its Undelayed Self") and Section 6.11.2.2 ("Partial-Response Filtering using a Comb Filter that adds Baseband DTV Signal to Itself as Delayed by Six Symbol Epochs"). "Precipitous" horizontal edges usually occur at the beginning and conclusion of every vertical sync interval in the interfering co-channel NTSC signal.

6.11.2.4 Partial-Response Filtering of DTV Signal using Comb Filtering with Differential Delays of Several Thousand Samples

Partial-response filtering of DTV signal using 262-NTSC-lines comb filtering subtracts the baseband DTV signal as delayed by 179,208 symbol epochs from its undelayed self.

Partial-response filtering of DTV signal using twice-NTSC-frame comb filtering subtracts the baseband DTV signal as delayed by 718,200 symbol epochs from its undelayed self. This the equivalent of a cascade connection of two NTSC-frame comb filters in cascade, one subtracting the baseband DTV signal as delayed by 359,100 symbol epochs from its undelayed self, and the other adding the baseband DTV signal as delayed by 359,100 symbol epochs to its undelayed self.[109]

The problem with these types of partial-response filtering is that a jump-cut, cut-through-black or lap dissolve in video program can cause demodulation artifacts of an interfering co-channel NTSC signal to be poorly suppressed for an entire field or frame of the NTSC signal. Demodulation artifacts are poorly suppressed in portions of the image with motion. The symbol decoding errors are then likely to be of such long duration that they cannot be corrected by the Reed-Solomon error-correction circuitry.

6.11.3 Partial-Response Filtering of Intermediate-Frequency DTV Signal to Suppress Demodulation Artifacts of Co-Channel NTSC Interference

Comb filtering to suppress demodulation artifacts of co-channel NTSC interference can be done in the IF passband, resulting in a partial-response intermediate-frequency DTV signal. A trellis decoder demodulates the partial-response signal, taking into account the states of the trellis encoder and the-partial response channel. The trellis decoder may have a 16-state trellis comprised of four 4-state butterflies wherein each edge in the trellis is a single transition.[110]

6.11.4 Suppression of Components of Co-Channel NTSC Interference by Selective Filtering

Selective filtering can be used to reject narrowband portions of the co-channel NTSC interference signal even though they overlap the DTV signal. For example, the video carrier can

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[109] K. S. Kim and M. K. Lee: "Cochannel Interference Filter HDTV Transmission System", U. S. patent No. 5 408 262, 18 April 1995.

[110] D. Rhee: "Trellis Code Modulation Decoder Structure For Advanced Digital Television Receiver", U. S. patent No. 6 201 563, 13 March 2001.

be suppressed, to remove a principal portion of continuing co-channel NTSC interference energy. The general approach is to isolate portions of the co-channel NTSC interference signal that have greater energy and then combine them destructively with the received signal. This selective-filtering approach is taken to reduce the 3 dB penalty in SNR associated with partial-response filtering.

Selective filtering is preferably done in a manner that not only reduces NTSC co-channel interference at specific frequencies, but also does not reduce SNR at those frequencies. Otherwise, the remnant noise, as referred to the time domain, exerts undesirable influence on adaptive equalization that will compromise suppression of multipath to some extent.

In specific embodiments of the selective-filtering approach, portions of the co-channel NTSC interference signal have been separated from a complex digital IF DTV signal by digital filtering and then suppressed using recursive filtering.[111] [112] [113] A practical problem with employing the selective-filtering approach at IF is that some NTSC signals are broadcast with offsets of their carrier frequencies from their nominal frequencies within channels. So, the selective-filtering approach is usually practiced at baseband.

In other specific embodiments of the selective-filtering approach, demodulation artifacts from portions of the co-channel NTSC interference signal have been separated from a complex digital baseband DTV signal by digital filtering and then suppressed using recursive filtering.[114] [115] [116] [117]

In other specific embodiments of the selective-filtering approach, demodulation artifacts from portions of the co-channel NTSC interference signal have been separated from a complex digital baseband DTV signal by digital filtering and then suppressed using a linear-phase causal FIR bandpass digital filter.[118]

The foregoing selective-filtering approaches have the problem that the portions of the co-channel NTSC interference signal that can be suppressed must be narrowband in nature, or the DTV signal is corrupted to the extent it cannot be satisfactorily detected. Many DTV receiver designers favor another species of the selective-filtering approach for suppressing demodulation artifacts of an interfering co-channel NTSC signal, which species utilizes the fact that these artifacts are less random in nature than the DTV symbols are. Besides the decision-feedback adaptive equalizer, there is an auxiliary adaptive filter that tunes itself for selecting periodic artifacts of the interfering co-channel NTSC signal from the baseband DTV signal before equalization. These periodic artifacts are fed forward to prevent their appearance in the response

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[111] C. G. Scarpa: "Apparatus for NTSC Signal Interference Cancellation Through the Use of Digital Recursive Notch Filters", U. S. patent No. 5 325 188, 28 June 1994.

[112] C. G. Scarpa: "Narrowband Interference Cancellation Through the Use of Digital Recursive Notch Filters", U. S. patent No. 5 325 204, 28 June 1994.

[113] C. G. Scarpa: "Method and Apparatus for NTSC Signal Interference Cancellation Using Recursive Digital Notch Filters", U. S. patent No. 5 400 084, 21 March 1995.

[114] S. N. Hulyalkar: "Method and Apparatus for Combating Co-Channel NTSC Interference for Digital TV Transmission", U. S. patent No. 5 452 015, 19 September 1995.

[115] S. N. Hulyalkar: "Method and Apparatus for Combating Co-Channel NTSC Interference for Digital TV Transmission", U. S. patent No. 5 512 957, 30 April 1996.

[116] S. N. Hulyalkar: "Method And Apparatus For Combating Co-Channel NTSC Interference for Digital TV Transmission Having a Simplified Rejection Filter", U. S. patent No. 5 602 602, 11 February 1997.

[117] S. N. Hulyalkar and M. Ghosh: "Blind Equalizer Method and Apparatus for HDTV Transmission Using an NTSC Rejection Filter for Mitigating Co-Channel Interference", U. S. patent No. 5 841 484, 24 November 1998.

[118] C. H. Strolle and S. T. Jaffe: "Digital Television Signal Processing System Including a Co-Channel Rejection Filter", U. S. patent No. 5 550 596, 27 August 1997.

of the adaptive equalizer. This approach is reported to increase TOV about 1.5 dB, rather than the 3 dB increase of TOV with simple partial-response comb filtering.[119] [120] [121]

Presuming the co-channel NTSC interference signal has sufficient energy to be synchronously detected, the video carrier and the DSB portions of its AM sidebands can be suppressed by the following procedure. The 750 kHz in-phase and quadrature-phase baseband video signals resulting from complex demodulation of the DSB portions of the video AM each contain artifacts of DTV signal. The artifacts of DTV signal are essentially the only components of the quadrature-phase baseband video signal and are Hilbert filtered to generate separated artifacts of DTV signal similar to those contained in the in-phase baseband signal. These separated artifacts of DTV signal are used to suppress the artifacts of DTV signal contained in the in-phase baseband video signal. The baseband video signal is re-modulated to IF and used for canceling the NTSC video carrier and the DSB portions of its AM sidebands accompanying the DTV IF signal. The magnitude of the baseband video signal recovered in this manner can be measured for use in deciding whether or not co-channel interference is sufficiently large that it should be suppressed.

In a receiver designed for receiving both NTSC and ATSC TV signals, this rather complicated filtering procedure also implements co-channel DTV interference being partially suppressed when NTSC analog TV signals are selected for reception.

In another rather complicated filtering procedure, co-channel interference from the NTSC FM audio carrier can be canceled by combining a DTV IF signal with its image as suitably translated in frequency. This procedure is adapted from one that has been used in video tape recording. There is some aliasing of tail frequencies in the Nyquist slope in this procedure, which aliasing causes some ISI. Suppressing NTSC FM audio carrier by SAW filtering of IF signal is simpler and as effective.

6.11.5 Suppression of Co-Channel NTSC Interference with Initialization of Equalizer Parameters from DFS Signal

The initial values of the weighting coefficients of the equalization filter are computed based on the received DTV signal previous to its equalization. Comb filtering to suppress co-channel NTSC interference can be performed on demodulated signal before baseband equalization filtering. Then, the response of this filtering to received DFS signal can be de-convolved by the response of this filtering to the ideal DFS signal, in order to generate the *channel impulse response* (CIR) from which the initial equalizer parameters can be computed. The problem with this approach is that the channel frequency response cannot be estimated accurately near the carrier nulls where the comb filtering response is mostly noise. De-convolution is most easily performed in the frequency domain using term-by-term division of DFT terms, but the null DFT terms in the comb-filtered DFS signal introduce further lack of precision in the deconvolution procedure. In the time domain, noise errors in the pulse components of the CIR are increased, which adversely affects the accuracy of the computation of initial weighting coefficients computed from the CIR.

A better approach is to follow the digital filtering to suppress co-channel NTSC interference with another partial-response digital filter that restores the spectrum of the DTV signal. Then, the

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[119] J. J. Nicolas and J. S. Lim: “Rapid-Update Adaptive Channel-Equalization Filtering for Digital Radio Receivers, such as HDTV Receivers”, U. S. patent No. 5 648 987, 26 September 1995.

[120] M. Ghosh: “Receiver-Based Methods and Devices for Combating Co-Channel NTSC Interference in Digital Transmission”, U. S. patent No. 5 572 262, 5 November 1996.

[121] M. Ghosh: “Receiver-Based Methods and Devices for Combating Co-Channel NTSC Interference in Digital Transmission”, U. S. patent No. 5 577 692, 7 July 1998.

response of the cascaded filtering to the received DFS signal can be de-convolved by the ideal DFS signal to generate the CIR from which the initial equalizer parameters can be computed. This avoids division-by-zero problems in performing de-convolution by DFT, and the dividend terms of the DFT are less likely to be noise-only terms than they are in the approach described in the preceding paragraph.

Another approach is to suppress the co-channel NTSC interference accompanying the DFS signal by accumulating six (or a multiple of six) successive DFS signals, which tends to average down the co-channel interference owing to phase slippage between DTV carrier and NTSC video carrier.[122]

6.12 Circuitry for Detecting Co-Channel NTSC Interference

There is an increase in random noise accompanying the baseband DTV signal when co-channel NTSC interference filtering is used. Accordingly, it is desirable to perform co-channel NTSC interference filtering of the reception channel only when the presence of significant co-channel NTSC interference is detected, either directly or inferentially. In the absence of significant co-channel NTSC interference being detected, the co-channel NTSC interference filtering is bypassed.

6.12.1 Circuitry for Detecting Co-Channel NTSC Interference Inferentially by Comparing Cumulative Data-Slicing Error With and Without Filtering to Suppress Interference

Many types of circuitry for detecting co-channel NTSC interference inferentially, by comparing cumulative data-slicing error with and without filtering to suppress such interference, have been designed in addition to that described in [3, Figure 9.8].[123] [124] [125] [126]

6.12.2 Circuitry for Detecting Co-Channel NTSC Interference by Detecting the Presence of Video Carrier

The energy around the NTSC picture carrier is sampled. This is compared with sampled white noise energy between the NTSC picture and color carriers, after field combing the digital television signal to eliminate the effects of static signals. The comparison is used to determine whether or not the NTSC rejection filter should be inserted in the digital television signal path.[127] This method of detecting the presence of video carrier is susceptible to error under certain multipath reception condition.

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[122] A. L. R. Limberg: "Suppression of Effects of Co-Channel NTSC Interference Artifacts upon Digital TV Receiver Adaptive Equalizer", U. S. patent No. 6 441 843, 27 August 2002.

[123] K. S. Kim and J. K. Kim: "NTSC Interference Detector", U. S. patent No. 5 546 132, 13 August 1996.

[124] L. E. Nielsen and G. J. Sgrignoli: "Detection of Co-Channel Interference in Digital Television Signals", U. S. patent No. 5 594 496, 14 January 1997.

[125] M. H. Lee: "Apparatus And Method Of Controlling Driving Selection Of NTSC Rejection Filter", U. S. patent No. 5 793 417, 11 April 1998.

[126] M. H. Lee: "Apparatus and Method for Canceling Co-Channel Interference", U. S. patent No. 5 969 751, 19 October 1999.

[127] R. W. Citta, L. E. Nielsen, and G. J. Sgrignoli: "NTSC Co-Channel Interference Reduction System", U. S. patent No. 5 821 988, 13 October 1998.

6.12.3 Intercarrier Presence Detector

Continuing presence of 4.5 MHz intercarrier in the received DTV signal can be detected for indicating co-channel NTSC interference.[128] This better discriminates against other forms of co-channel interference than detecting the presence of video carrier. Also, the intercarrier presence detector is less apt to be affected by multipath distortion.

6.12.4 Circuitry for Detecting Co-Channel NTSC Interference in Auxiliary NTSC Analog TV Receiver that Suppresses Co-Channel DTV Signal

An auxiliary NTSC analog TV receiver provides full-band synchronous detection for imaginary signal as well as real signal. The imaginary signal is inverse-Hilbert-transformed and combined with the real signal to separate NTSC baseband video signals up to 750 kHz from co-channel DTV interference. The level of this detected NTSC baseband video signal, if any, is measured to determine the severity of NTSC co-channel interference.[129]

6.13 Channel Equalizer

6.13.1 Methods for Initializing Time-Domain Equalizer Parameters

The multipath reception conditions encountered in over-the-air DTV broadcasting using transmission antennas mounted on ground-based towers mandate that the equalizers in DTV receivers for receiving such broadcasting be adaptive in nature. The fairly high probability at many reception sites of appreciable change in multipath conditions every few milliseconds makes it desirable that adaptation of the equalizer is performed more frequently than once every data field. So, the data-based adaptation of equalizer coefficients by a Kalman-type auto-regression technique is used in most designs of DTV receivers for the consumer market. The initialization of equalizer filtering parameters by such a technique tends to be very slow until some tracking of the adaptation to changing multipath conditions can be established and indeed initialization may never be attained in a short enough time to be practical. So, initialization of equalizer filtering parameters using at least one other technique is customary in designs for DTV receivers for the consumer market. Initialization of equalizer filtering parameters within a fraction of a second is desirable in a DTV receiver designed for the consumer market.

6.13.1.1 Measuring Channel Impulse Response (CIR) using the Received Triple-PN63 Sequence

The PN63 sequences in the DFS signal are maximal-length and were chosen for flatness of frequency spectrum. So, CIR can be extracted from a portion of the response of a match filter having a PN63 kernel, used as a correlation filter. The weighting coefficients for the receiver equalizer can then be calculated from measured CIR using the procedures described in Section 5.12.1 (“Computation of Initial Weighting Coefficients of the Adaptive Equalizer Based on CIR Measurements”), *supra*.

Because of the PN63 sequence being repeated and then repeated again in the received DTV signal, the convolution of one PN63 sequence with three consecutive PN63 sequences of the same polarity in certain of the received DFS signals generates a correlation result having three equal peaks at 63-symbol spacing. The regions of the correlation between the central peak and each of the outside peaks is a constant equal to –1/63 of the peak value, so auto-correlation noise in these regions is almost 36 dB down. To the extent that there is low correlation noise from

[128] A. L. R Limberg: “Using Intercarrier Signals for Detecting NTSC Interference in Digital TV Receivers”, U. S. patent No. 5 923 378, 13 July 1999.

[129] A. L. R. Limberg: “Using Special NTSC Receiver to Detect When Co-Channel Interfering NTSC Signal Accompanies a Digital TV Signal”, U. S. patent No. 5 852 476, 22 December 1998.

echoes that overlap the triple-PN63 sequence and arise from signal that precedes or succeeds the triple-PN63 sequence, a reasonably accurate CIR measurement extending over a 63-symbol-epoch interval can be extracted directly from the PN63 match filter response. The preceding PN511 sequence exhibits a considerable degree of orthogonality to the PN63 sequence in all their possible respective phases, so PN63 match filtering response to even strong post-echoes of the PN511 sequence will be relatively low compared to the central and outside peaks. Post-echoes of symbols preceding the PN511 sequence that overlap the triple-PN63 sequence will usually be weak, since this is generally the case with long-delayed echoes. The principal concern with regard to correlation noise is considerably advanced pre-echoes of the symbols succeeding the triple-PN63 sequence happening to correlate reasonably well with the PN63 sequence.

The receiver designer may arbitrarily apportion the total 63-symbol echo-suppression range into respective ranges for pre-echoes and for post-echoes. This is done by choosing the phase of the PN63 sequence that is used for match filtering. Two PN63 match filters with suitably phased kernels can be used, one to measure a 63-symbol-epoch pre-echo range, and the other to measure a 63-symbol-epoch post-echo range. These CIR measurements are combined to provide a CIR extending over a wider echo range. Weighting coefficients for the receiver equalizer can then be calculated from this extended CIR using the procedures described in Section 5.12.1 ("Computation of Initial Weighting Coefficients of the Adaptive Equalizer Based on CIR Measurements"), *supra*.

PN63 match filters with kernels of slightly different phasing can be used to generate respective CIR measurements with different correlation noise characteristics. A synthetic CIR measurement can then be synthesized from the components of the individual CIR measurements least likely to be affected by correlation noise.

The triple-PN63 sequences can be processed in another way for equalizing DTV signals in which multipath reception conditions do not change too much between DFS signals. Differentially combining the DFS signals from two successive data fields will result in a single baud-rate PN63 sequence preceded by a 515-symbol-epoch zero-value interval and succeeded by another zero-value interval of at least 63 symbol epochs duration, providing that reception is not multipath in nature. In practice there is almost always multipath, causing the single baud-rate PN63 sequence to be accompanied by echoes thereof. The 578 symbols preceding this single baud-rate PN63 sequence will still cancel, along with their echoes, providing that multipath conditions do not change appreciably from the earlier DFS signal to the later one. At least 63 symbols succeeding this single baud-rate PN63 sequence will still cancel, along with their echoes providing that multipath conditions do not change appreciably from the earlier DFS signal to the later one. If the 24-symbol VSB mode code and 92 of the following reserved symbols do not differ in the two DFS signals being differentially combined, 179 symbols succeeding this single baud-rate PN63 sequence will still cancel, along with their echoes providing that multipath conditions do not change appreciably from the earlier DFS signal to the later one. The response of a PN63 match filter to the isolated baud-rate PN63 sequence and its echoes can be used to generate a CIR measurement. This CIR measurement method can be used on accumulation results for a number of pairs of successive DFS signals, so correlation noise from symbols following the triple-PN63 sequence is averaged and hopefully so reduced.

6.13.1.2 Measuring Channel Impulse Response using the Received PN511 Sequence

The PN511 sequence in the DFS signal is designed to have a uniform spectral response. While CIR measurement can be done by de-convolving the received PN511 sequence with echo-free PN511 sequence, the measurement of CIR by correlating the received PN511 sequence with echo-free PN511 sequence is less apt to exhibit noise arising from quantization effects. If isolated and not wrapped around, the PN511 sequence in the DFS signal exhibits an auto-correlation function that has a single central peak surrounded by small values at least 30 dB lower than the central peak. This residual correlation noise limits the accuracy with which the amplitudes of multipath components can be measured to approximately 27 dB below the principal signal. Because the PN511 sequence in the DFS signal is not isolated from surrounding signal, it also exhibits random correlation to adjacent parts of the transmitted signal. On average, this random correlation has a level of approximately –27 dB compared to the peak correlation for the principal signal. At times the correlation of the PN511 signal with an adjacent part of the transmitted signal or a strong echo thereof can be significantly higher than –27 dB, however, causing considerable error in echo measurement. The measurement of multipath components as small as 27 dB below the principal signal suffices for adequately de-echoing an 8VSB signal, so long as the number of echoes is not too large.

The random correlation noise can be further reduced by various means that effectively average several DFS signals. Such averaging can reduce the effect of the occasional higher-than-normal correlation of the PN511 signal with an adjacent part of the transmitted signal or a strong echo thereof. Furthermore, the averaging procedures can be designed to discount the effect of occasional anomalous echo measurements.

6.13.1.3 Measuring Channel Impulse Response using the Received String of PN 511 Sequence and Triple PN63 Sequence

CIR measurement can be done by de-convolving the received string of DSS sequence, PN 511 sequence and triple PN63 sequence with the echo-free string as stored at the receiver. The 704-symbol string of DSS sequence, PN 511 sequence, and triple PN63 sequence extends over more symbol epochs than the PN511 sequence alone. So, the edge effects caused by the 704-symbol string being overlapped by echoes of symbols that precede or succeed the string have less influence on the de-convolution result. De-convolution is usually done using a *discrete Fourier transform* (DFT) method.

Alternatively, the received strings of DSS sequence, PN 511 sequence and triple PN63 sequence from the initial data segments of two successive data fields can be combined, and an auto-correlation procedure can be used for the CIR measurement. The two 704-symbol strings are combined to suppress inaccuracies in the CIR measurement caused by the PN sequence being repetitive. The echo-free string stored at the receiver for use in the auto-correlation is then that which would result from combining the echo-free strings of DSS sequence, PN 511 sequence and triple PN63 sequence from the initial data segments of two successive data fields.

6.13.1.4 Measuring Channel Impulse Response (CIR) using Long Data Sequences

The data in the 8VSB signal are randomized, so there is marked symbol variation in the 8-VSB signal. Accordingly, self-correlation of a long string of 4096 symbols or more generates a complex-CIR measurement that is little affected by edge effects caused by the long symbol string being overlapped by echoes of symbols that precede or succeed it. While the amplitudes of the complex-CIR components are quite accurate, the phases of the complex-CIR components are not as precisely defined in time as those measured using the 704-symbol string of DSS sequence, PN 511 sequence, and triple PN63 sequence. Accordingly, the amplitudes of the complex-CIR components can be measured using a long string of 4096 symbols immediately preceding the 704-symbol string of DSS sequence, PN 511 sequence, and triple PN63 sequence used in measuring the phases of the complex-CIR components. This provides a good initial complex-CIR measurement for adaptive equalizers that update that measurement response to reception error measurements of equalizer response and then calculate updated weighting coefficients for the equalizer from the updated complex-CIR measurement.

6.13.1.5 Initialization of Equalizer Parameters Based on Previous Reception of a Channel

Memory addressed by channel number can be used to store the equalizer parameters for a reception channel when that channel is no longer selected for reception or when the DTV receiver is no longer fully powered. When the channel is next selected for reception or the DTV receiver is again fully powered, the equalizer parameters can be restored based on read-out from the memory.[130] This increases the likelihood of re-establishing tracking of the adaptation of equalizer filtering parameters by a data-based technique.

Alternatively, memory addressed by channel number can be used to store information for regenerating the equalizer parameters for a reception channel when that channel is no longer selected for reception or when the DTV receiver is no longer fully powered, rather than storing the equalizer parameters *per se*.[131] The stored information may be the time-domain channel impulse response (CIR), or some portion thereof, by way of examples.

6.13.1.6 Initialization of Adaptive Equalizer Coefficients using Constant Modulus Algorithm (CMA)

CMA is a special case of the "Godard algorithm" which permits equalizer adaptation that, although it is quite slow, does not require carrier phase recovery.[132] [133] CMA was developed for signals with both upper and lower sideband structures, such as PSK and QAM. CMA relies on the fact that, at the decision instants, the modulus of the detected data symbols should lie on one of several circles of varying diameters on the two-dimensional constellation plane. Like the decision-directed LMS algorithm, the CMA algorithm is usually implemented with a gradient descent strategy in which the equalizer parameters are adapted by replacing the present equalizer parameter settings with their current values plus an error (or correction) term.[134] The CMA error term itself is a cubic function of the equalizer output.

If the symbol timing of the VSB signal is substantially exact, it may be converted into a *staggered-quadrature-amplitude-modulated* (SQAM) signal by shifting its carrier frequency by $f_{(S/4)}$, where f_S is the symbol frequency, and the SQAM signal may be converted to a QAM signal by delaying the I samples by one-half symbol period. This conversion is based on the known fact that a SQAM signal having an offset equal to T/2, where T is the baud interval, is equivalent to a VSB signal with a carrier frequency $\omega_0 \pm(\pi/2)$ and with identical pulse and vestigial spectral shaping.[135] This QAM signal can be demodulated to provide a complex baseband signal suited to application of the CMA algorithm for determining the weighting coefficients of a complex baseband equalizer.[136] Alternatively, the weighting coefficients of a complex passband equalizer can be determined by accounting for the effects of de-rotation during demodulation.[136] A form of

130 D. J. Kim: "Channel Equalizer for Digital Television Receiver Having an Initial Coefficient Storage Unit", U. S. patent No. 5 654 765, 5 August 1997.

131 C. B Patel: "TV Receiver Equalizer Storing Channel Characterizations for Each TV Channel Between Times of Reception There From", U. S. patent No. 6 222 592, 24 April 2001.

132 D. N. Godard: "Self-recovering Equalization and Carrier Tracking in Two-dimensional Data Communication Systems," *IEEE Trans. on Communications*, vol. 28, no 11, pp. 1867–1875, October 1980.

133 J. R. Treichler and B. G. Agee: "A New Approach to Multi-Path Correction of Constant Modulus Signals," *IEEE Trans. On Acoustics, Speech and Signal Processing*, vol. ASSP-31, no.2, pp. 459–472, April 1983.

134 C. R. Johnson, Jr., C. R. Schniter, T. J. Endres, J. D. Behm, D. R. Brown, and R. A. Casas: "Blind Equalization Using the Constant Modulus Criterion: A Review," *Proc. of the IEEE*, vol. 86, no. 10, pp. 1927–1950, October, 1998.

135 R. Gitlin, et al.: "The Performance of Staggered Quadrature Amplitude Modulation in the Presence of Phase Jitter," *IEEE Transactions on Communication*, vol. com-23, no. 3, pp 348–352, March 1975.

136 T. J. Endres, S. Hulyalkar, T. A. Schaffer, and C. H. Strolle: "Reduced Complexity Equalizer for Multi Mode Signaling", U. S. patent No. 6 426 972, 30 July 2002.

CMA called Single-Axis CMA (SA-CMA) can be used to constrain the weighting coefficients of the baseband equalizer or the passband equalizer to be real.[136 137]

The CMA algorithm can be applied directly to a complex VSB signal, so long as the complex VSB signal is not synthesized by phase-splitting a real-only VSB signal that is sampled at less than twice Nyquist rate. Sampling a real-only VSB signal at less than twice Nyquist rate under-samples the modulus of the detected data symbols within the two-dimensional constellation plane, causing phase ambiguities that give rise to local minima in the CMA cost function. The CMA algorithm can stall in these local minima, which are undesirably closed-eye.[138]

Quantizing the received signal reduces the number of values used in the CMA error calculation, so a relatively small lookup table can be used to compute the CMA error function for VSB signals.[139] This computation can be done directly on a VSB signal without need for converting the VSB signal to a QAM signal. CMA has been applied to fractional equalizers.[140] In some equalizer designs different quantization levels are used in different regions of the CMA error function.[141]

6.13.1.7 Initialization of Adaptive Equalizer Coefficients using Modified Reduced Constellation Algorithm (MRCA)

The reduced constellation algorithm (RCA) relies on forming "super constellations" within the main constellation. The data signal is first forced to fit into a "super constellation," and then the "super constellations" are subdivided to include the entire constellation. As conventionally used, RCA inherently relies on a two-dimensional constellation being used. However, blind equalization can be performed on the VSB constellation using a modified reduced constellation algorithm (MRCA). The key part of this modification is to realize that there exists a one-dimensional version of the RCA algorithm which is appropriate for VSB signals. The MRCA consists of an algorithm to determine appropriate decision regions for a VSB decision device, so as to generate decisions that allow an adaptive equalizer to converge without the use of a training signal.

In VSB systems, the decision regions typically span one data symbol of the full constellation and the upper and lower bounds of each decision region are set midway between the constellation points. If these decision regions are used for initial convergence of the equalizer, the equalizer will not converge, since—due to the presence of intersymbol interference—a significant amount of the decisions from the decision device will be incorrect.

In order to force more correct decisions to be made, an algorithm for determining new upper and lower decision region boundaries has been determined. The algorithm clusters the full VSB constellation into several sets, determines upper and lower bounds for decision regions, and appropriate decision device output "symbol" values. These first sets are further divided into smaller sets until each set of symbols contains exactly one symbol, and the decision regions

137 A. Shah, S. Biracre, R. A. Casas, T. J. Endres, S. Hulyalkar, T. A. Schaffer, and C. H. Strolle: "Global Convergence of a Single-Axis Constant Modulus Algorithm," IEEE Statistical Signal and Array Processing Workshop, Pocono Manor, PA, pp. 645–649, August 2000.

138 C. B. Papadias: "On the Existence of Undesirable Global Minima of Godard Equalizers', International Conf. on Acoustics, Speech and Signal Proc., Munich, Germany, pp.3941–3944, April 21–24, 1997.

139 T. J. Endres, S. N. Hulyalkar, and C. H. Strolle: "Adaptive Equalizer with Decision Directed Constant Modulus Algorithm", U. S. patent No. 6 337 878, 26 September 2000.

140 I. Fijalkow, A. Touzni, and J. R. Treichler: "Fractionally spaced equalization using CMA: Robustness to channel noise and lack of disparity," *IEEE Trans. on Signal Processing*, vol. 45, no. 1, pp.55–66, January 1997.

141 T. J. Endres, S. Hulyalkar, C. H. Strolle, and T. A. Schaffer: "Adaptive Equalizer with Enhanced Error Quantization", U. S. patent No. 6 418 164, 9 July 2002.

correspond to the standard decision regions for VSB described above. The function of each stage is to allow for more decisions to be correct, and thereby drive the equalizer towards convergence. In this way, each stage in the blind equalization process serves to further open the eye.

In general, the MRCA algorithm consists of clustering the decision regions of the decision device into finer portions of the VSB constellation. The method starts with a binary (two-level) slicer, then switches to a four-level slicer, then eight-level slicer, etc.

Note that the MRCA algorithm is applicable both to linear equalization and to decision feedback equalization.

6.13.2 Switch-Over from Initialization to Data-Directed Adaptation

There are procedures for determining when equalization filtering parameters should be initialized or re-initialized, and when equalization filtering parameters should be updated by a data-directed method. Pronounced low-frequency variation (or "flutter") in the baseband DTV signal can be detected and used as an indication that data-directed methods of adaptation are apt to experience a tracking failure. The gain factor μ controlling the increment of decision error used in adaptation can be reduced when flutter is detected. Another approach compares the most current equalized DFS with the recent average of equalized DFS and institutes initialization if the comparison is not good enough.[142] [143]

6.13.3 Cross-Over Between FIR and IIR Filtering in the Equalizer

As noted in Section 5.4 ("Equalizer Filter Implementation in the Time Domain"), the manner in which the cross-over between FIR filtering and IIR filtering is made can significantly affect equalizer performance when the principal signal is accompanied by a short-delay echo having substantial energy or by a larger number of such echoes. In some DTV receiver designs, the feedback FIR filter is used exclusively for processing decision feedback from the Viterbi trellis decoding. The equalizer response in the time domain overlaps the response of the feed-forward FIR filter and the response of the feedback FIR filter used for IIR filtering, rather than abutting such responses. The responses may overlap more than a hundred symbol epochs. The need for overlaps this large arises because the feed-forward FIR filter is called upon to suppress short-delay post-echoes, owing to the delay in the trellis decoding providing decision feedback to the feedback FIR filter used for IIR filtering. The suppression of short-delay post-echoes by the FIR filter generates continuing repeats, and those of which have significant energy are also suppressed by the feed-forward FIR filter. There can be several cycles of repetition if the energy of any original short-delay post-echo is nearly as great as the energy of the principal signal. If inadequate provision is made for suppressing the repeats of short-delay echoes having appreciable energy, the DFE response will contain appreciable amounts of ISI.

A simple and effective approach for reducing the need for the feed-forward FIR filter response to overlap the response of the IIR filtering with decision feedback is to suppress short-delay echoes using IIR filtering with linear feedback, rather than suppressing the short-delay echoes with the feed-forward FIR filter. Short-delay echoes are suppressed by IIR filtering with linear feedback without generating repeats.

Parallel decision-feedback equalization can suppress short-delay post echoes, substantially reducing the need for the feed-forward FIR filter response to overlap the response of the IIR

[142] R. W. Citta and L. E. Nielsen: "System for Controlling the Operating Mode of an Adaptive Equalizer", U. S. patent No. 5 572 547, 5 November 1996.

[143] L. E. Nielsen: "System for Controlling the Operating Mode of an Adaptive Equalizer", U. S. patent No. 5 684 827, 4 November 1997.

filtering with parallel decision feedback. Another approach for reducing the need for overlap is to reduce the amount of time to obtain decisions from the Viterbi trellis decoding, so there are fewer and shorter-delay post-echoes that the feed-forward FIR filter has to suppress. “Smart” data slicers exploit properties of the trellis coding to supply decisions in real time.[144]

6.13.4 Block Computation of Equalizer Coefficients from Continuously Updated CIR

The auto-regression techniques that have been used for updating the weighting coefficients of Kalman equalization filters can be applied to updating the CIR instead. The weighting coefficients of the equalization filter can then be periodically computed from a strobe of the then-current CIR for a simultaneous block update of all the weighting coefficients. The principal component of the strobed CIR is chosen, and the components of the strobed CIR are normalized respective to that chosen principal component.

The portion of the normalized strobed CIR corresponding to the longer-delay post echoes that can be canceled by IIR filtering using decision feedback is identified. Because the feedback FIR filter cancels echoes using variously-delayed equalizer response and echoes are suppressed in the equalizer response, the coefficients of the feedback FIR filter are the complements of corresponding terms of that portion of the normalized strobed CIR.

The feed-forward FIR filter suppresses echoes using variously-delayed equalizer input signal, which equalizer input signal has accompanying echoes. So, the normalized strobed CIR is processed to replace the principal component and longer-delay post-echoes there from and then is de-convolved by the complete normalized strobed CIR to determine the echo structure that must be suppressed by variously-delayed equalizer input signal that has echoes. This de-convolution procedure takes into account the convolution of the FIR filter kernel with the input signal supplied to the equalizer, which input signal comprises a principal component that can be accompanied by one or more echoes with significant individual energy, rather than being echo-free. The echo structure generated by this de-convolution contains all the repeats that can be resolved by the feed-forward FIR filter for the digital sampling resolution being used. The coefficients of the feed-forward FIR filter are the complements of corresponding terms of that echo structure.

De-convolution is most readily performed using DFT. The apparatus for implementing this is the same as that used for initialization from DFS signal using the Dietrich-Greenberg method, as described in Section 5.12.1 (“Computation of Initial Weighting Coefficients of the Adaptive Equalizer Based on CIR Measurements”). The coefficients of the feed-forward FIR filter can easily be generated within the duration of a data segment or so, during which time CIR updates are accumulated. This compares favorably with other auto-regression methods that update weighting coefficients in block, but do not do so depending on updating of CIR. The intervals at which block updating is done can be as long as 3328 symbol epochs (4 data segments), but must be considerably shorter if fast-changing multipath distortion is to be accommodated.

The echo structure to be suppressed by the feed-forward FIR filter can be superposed on the long-delayed echo portions of the normalized strobed CIR to generate a synthetic normalized strobed CIR. The echo terms of this synthetic normalized strobed CIR that are not so long delayed that decision feedback is unavailable can be suppressed substantially in their entirety by the IIR filtering, rather than by the feed-forward FIR filter. This reduces the overlap that the feed-forward FIR filter must have with the IIR filtering and can save a considerable amount of digital multiplication. The echo terms of the synthetic normalized strobed CIR that are not so

[144] S. N. Hulyalkar, T. J. Endres, T. A. Schaffer, and C. H. Strolle: “Method of Estimating Trellis Encoded Symbols Utilizing Simplified Trellis Decoding”, U. S. patent No. 6 178 209, 19 June 1998.

long delayed that decision feedback is unavailable are complemented to determine the coefficients of the feedback FIR filter. Suppressing these terms using decision feedback helps control noise growth in the equalization process.

Using auto-regression methods to update the CIR, rather than to update the weighting coefficients of the equalization filter directly, allows accurate computation of filter coefficients of the FIR filtering portion of the equalization filter to be done more quickly. Pre-echoes and short-delay echoes are suppressed in the feed-forward FIR filter portion of the equalization filter by combining them with suitably delayed equalization filter input signal. Equalizers using Kalman-type filters depend on the equalization filter input signal having a principal component substantially stronger than echo components. This is so the filter parameters calculated for suppressing echoes with an echo-free signal approximate reasonably well the filter parameters for suppressing echoes with the equalization filter input signal, which is apt to include echoes. When the received signal has severe multipath distortion, the filter parameters calculated by auto-regression techniques for suppressing echoes with an echo-free signal are not a very good approximation of the filter parameters for suppressing echoes with the equalization filter input signal. An adaptive equalizer using a Kalman-type filter will take quite a long time for its weighting coefficients to converge to correct values. Indirect computation of the parameters of the equalization filter proceeding from the CIR facilitates taking into account the effects of the echoes in the equalization filter input signal used for suppressing echoes. There is a latency in calculating the equalization filter parameters from the CIR, but there is no convergence procedure that requires time to complete.

The ready availability of the normalized strobed CIR provides a basis for making decisions concerning the use of spectrum combining, partial-equalization, sparse equalization, or other special equalization techniques.

6.13.5 Partial-Response Equalization to Reduce Noise Growth in Reception Channels having Spectral Nulls and/or Near-Nulls

Attempting to flatten the spectral response of a channel exhibiting spectral nulls and/or near-nulls results in undesirable noise growth in the low-energy portions of the spectrum. A better procedure is to apply the received signal to a noise-predictive partial-response equalizer consisting of a linear equalizer, which shapes the channel response to a predetermined partial-response function, followed by a linear predictor such as a partial-response Viterbi trellis decoder. This scheme modifies the output sequence of the linear partial-response equalizer by whitening the total distortion; i.e. by whitening the noise components and the residual interference components at the equalizer output, thereby achieving the best possible signal-to-noise ratio (SNR) before detection.[145] [146] The particular type of linear partial-response required from the linear equalizer can be determined from the CIR measured during DFS interval(s).

6.13.6 Passband Equalization

Most DTV designers prefer baseband decision-feedback equalization, but passband decision-feedback equalization has been considered.[147] A VSB modulator has to be included in the decision-feedback loop if passband equalization is used. The latent delay through the VSB

[145] P. Chevillat, E. Eleftheriou, and D. Maiwald: "Adaptive Noise-Predictive Partial-Response Equalization for Channels with Spectral Nulls", U. S. patent No. 5 784 415, 21 July 1998.

[146] P. Chevillat and E. Eleftheriou: "Fully Adaptive Modem Receiver Using Whitening Matched Filtering", U. S. patent No. 5 031 195, 9 July 1991.

[147] C. B. Patel and A. L. R. Limberg: "Digital Television Receiver with Equalization Performed on Digital Intermediate-Frequency Signals", U. S. patent No. 6 124 898, 26 September 2000.

modulator causes the decision-feedback loop in the IIR filtering not to be able to accommodate some less-delayed post-echoes. So, these less-delayed post-echoes have to be suppressed by FIR filtering.

Passband equalization may be used just in the feed-forward portion of the equalization filtering, with decision-feedback equalization being done at baseband.[148]

Passband equalization also makes it difficult to compute initial weighting coefficients for the equalizer based on CIR determined by the receiver from comparing the received DFS signal with the ideal DFS signal stored in the receiver. The received DFS signal must be de-rotated before equalization, so it can be compared with the ideal DFS signal stored in the receiver to determine CIR. This de-rotation before equalization is in addition to the de-rotation after equalization, which latter de-rotation is done to recover equalized baseband DTV signal for data-slicing. There is also the complication of having to compute complex passband equalization filter coefficients from the CIR, which computation involves a rotation procedure reversing the de-rotation procedure before equalization.

6.13.7 Frequency-Domain Equalization

6.13.7.1 Frequency-Domain Equalization in the Digital Regime

Equalization can be performed in the frequency domain, rather than in the time domain. Some designers have investigated the use of frequency-domain equalizers. However, all DTV receivers designed for the consumer market that were sold in the 1996-2002 timeframe use time-domain equalizers, most if not all of these designs using baseband rather than passband equalizers. Noise growth during equalization of post-echoes is controlled quite well by decision-feedback methods in time-domain processing, and the desire to benefit from decision feedback is probably the best reason for favoring time-domain equalizers over frequency-domain equalizers. Also, fewer operations appear to be involved in time-domain processing.

In frequency-domain equalizers the discrete Fourier transform of the DTV signal is obtained and multiplied, term by term, by the DFT of a function that would provide equalization in the time domain. The *inverse discrete Fourier transform* (I-DFT) of the resulting DFT generates the DTV signal as it would be received over an ideal reception channel.

6.13.7.2 Frequency-Domain Equalization in the Analog Regime

Some multipath conditions reduce the energy in certain parts of the channel frequency spectrum so severely that successful reception is impossible unless the IF passband is altered. This alteration should be done before digitization, because one of the reasons for unsuccessful reception is quantization noise introduced during A/D conversion obscuring or obliterating low-spectral-energy components of the DTV signal. The need for this type of compensation can be determined from the DFT of the channel impulse response (CIR) or, alternatively, from the CIR itself assuming certain proscribed patterns. Generally, the best of only a few alternative IF passbands will be selected for reception. So, frequency-domain equalization in the analog regime will usually be augmented by frequency-domain equalization in the digital regime.

6.13.8 Bibliography of Documents Regarding Equalization but not Specifically Cited

H. H. Zeng, L. Tong, and C. R. Johnson, Jr.: “An Analysis of Constant Modulus Receivers,” *IEEE Trans. on Signal Processing*, vol .47, no.11, pp. 2990–2999, November 1999.

[148] C. H. Strolle and S. T. Jaffe: “Blind Equalizer Method And Apparatus For HDTV Transmission Using an NTSC Rejection Filter for Mitigating Co-Channel Interference”, U. S. patent No. 5 550 596, 24 November 1998.

P. Schniter and C. R. Johnson, Jr.: "Bound for the MSE performance of Constant Modulus Estimators," *IEEE Trans. on Info. Theory*, vol. 46, no. 7, pp. 2544–2560, November 2000.

6.14 Treatment of Zero-Frequency Pilot Accompanying Baseband DTV Signal

The real baseband DTV signal recovered by demodulating an intermediate-frequency DTV signal will include a direct component corresponding to the pilot carrier having been synchrodyned to zero frequency. The amplitude of this direct component is nominally 1.25 times the spacing between modulation levels in the DTV signal, but is apt to be different because of multipath distortion.

The use of comb filtering to suppress NTSC interference cancels the direct component resulting from pilot detection. Measures should be taken to remove this direct component before data slicing when the comb filtering to suppress NTSC interference is not used, supposing that the same data slicing procedures are to be used as when comb filtering to suppress NTSC interference is used. Suppressing the direct component resulting from pilot detection before equalization filtering reduces the dynamic range required in its multipliers, so is preferred in designs that equalize only real baseband signal. The direct component is suppressed after equalization, however, in designs with complex equalization filtering that use the equalized imaginary baseband DTV signal for AFPC of the local oscillations used in synchronous detection of the intermediate-frequency DTV signal.

Suppressing the direct component resulting from pilot detection is complicated by the fact that the 8-VSB and 16-VSB signals are *non-return-to-zero* (NRZ) codes. So, despite randomization, these signals may occasionally include a long-lived transient direct component. Accordingly, simply highpass filtering the baseband DTV signal to suppress the direct component resulting from pilot detection will occasionally cause erroneous data slicing, because the long-lived transient direct components of data are also suppressed.

The IF signal can be periodically interrupted to develop a zero-carrier condition to establish a baseline against which to measure the direct component of the baseband DTV signal.[149]

The direct component of the baseband DTV signal is measured just during the DFS and DSS sequences in some procedures for suppressing the direct component resulting from pilot detection without suppressing the long-lived transient direct components of data. The symbols in these sequences are sorted according to prescribed polarity. The absolute values of the symbols in the two sets of sorting results are each averaged. The two averages are differentially combined and scaled to generate the measurement. The measurement is used to define a direct component offset subtracted from the baseband DTV signal in order to suppress the direct component resulting from pilot detection.

Alternatively, each successive pair of DFS-interval signals can be averaged to cancel the middle PN63 pulse signal and leave a pedestal 63 symbol epochs long that is the direct component resulting from pilot detection.

6.14.1 Bibliography of Documents Regarding Negating dc Offset in the Baseband DTV Signal but not Specifically Cited

R. B. Lee and L. E. Nielsen: "System for Negating the Effects of DC Offsets in an Adaptive Equalizer," U. S. patent No. 5 060 067, 22 October 1991.

D. J. Kim: "Device for Correcting DC of HDTV", U. S. patent No. 5 696 559, 9 December 1997.

R. Turner: "DC Removal Circuit for Digital Signal", U. S. patent No. 5 778 028, 7 July 1998.

149 G. J. Sgrignoli: "DC Offset Compensation Method and Apparatus", U. S. patent No. 5 669 011, 16 December 1997.

L. K. Tan, T. M. Liu, and H. A. T. Hung: "Technique for Minimizing Decision Feedback Equalizer Word Length in the Presence of a DC Component", U. S. patent No. 6 438 164, 20 August 2002.

6.15 Phase Tracker

6.15.1 Phase Tracking Before Decision-Feedback Equalization

In the Grand Alliance DTV receiver [3], the phase tracker is located after the baseband equalizer. In DTV receivers employing a baseband equalizer with decision feedback, it is preferable to locate the phase tracker (or symbol synchronizer) after the feed-forward FIR filter in the equalizer, but before the subsequent IIR filter.[150] The subsequent IIR filter includes a digital adder for combining the responses of the feed-forward FIR filter and a feedback FIR filter to generate a sum signal. This sum signal is re-sampled to symbol rate, if necessary, to provide the equalizer response supplied to the trellis decoder. The equalizer response is data sliced to generate a decision that is symbol-coded and supplied as decision-feedback input signal to the feedback FIR filter.

The phase tracker adjusts symbol phasing for minimum intersymbol interference (ISI) in the data-slicing used to generate the decision-feedback input signal for the feedback FIR filter. However, the phase tracker is outside the decision-feedback loop in the IIR filter, so the latent delay of the phase tracker is not included in the decision-feedback loop delay. This is desirable, so shorter-delay post-echoes can be suppressed by the IIR filter employing the decision feedback. The phase tracker aligns the entire IIR filtering operation with the optimum symbol phasing to minimize ISI in the data-slicing operation for generating decision-feedback signal.

The control signal for the phase tracker (or symbol synchronizer) is generated from the equalizer response. The subsequent digital adder for combining the responses of the feed-forward and feedback FIR filters introduces very little additional delay into the control loop.

6.16 Trellis Decoder

The term "trellis code" refers to a code generated by a finite-state machine, in which input of the information to be coded causes transitions or branches among states, each branch accompanied by the output of a group of encoded symbols. In a convolutional code, the finite-state machine is a shift register, and the output is produced by operations on the contents of various register stages.

The Viterbi method of trellis decoding is basically a maximum-likelihood method that maintains, in a path memory, a record of the sequence of branches, or path, by which each state is most likely to have been reached at the present time. Each time a new group of symbols is received, branch metrics are calculated for all possible branches from each state, and the most likely paths are updated on the basis of these branch metrics and existing path metrics. In most designs, the *add-compare-select* (ACS) operations used to generate these updates are performed by a number of ACS units in a digital signal processor (DSP) served by a branch metric memory and by the path memory, typically comprising a path metric memory and a path selection signal memory. At the end of a received message, one path is selected and retraced, and its branches are decoded to reconstruct the original data. One problem with the basic Viterbi method is that for long messages, a very large path memory is required to hold all the associated path information. Another problem is a long decoding delay, because decoding does not begin until the entire

[150] A. L. R. Limberg: "PAM Radio Signal Receiver with Phase-Tracker Succeeding Adaptive FIR Filtering and Preceding Adaptive IIR Filtering", U. S. patent application filed 27 January 2003.

message has been received. The conventional solution to both of these problems, known as path *memory truncation*, stores path information only for a certain number of most recent branches. Each time a new group of symbols is received, the most likely path is retraced back to its oldest stored branch; this single branch is decoded; then the oldest branches of all paths are deleted from the path memory to make space for new branches. Straightforward path memory truncation exacts a heavy processing-load penalty, because the path retracing process must be carried out every time a new group of symbols is received. Some receiver designs do not use 12-phase trellis decoding, but instead use single-phase Viterbi decoding.[151] Two-state state metrics speed Viterbi decoding.[152] [153] [154]

A considerable amount of effort has gone into trimming down the hardware and software required for Viterbi trellis coding. Less memory is needed for storing path information if it is arranged that the most recently traced paths overwrite the least recently traced paths after decoding the information contained therein.[155]

Use of a plurality of add-compare-select units speeds up decoding in a Viterbi decoder. The Viterbi decoder requires memory for reading state metrics to these ACS units and for writing state metrics from these ACS units. The storage capacity for storing the state metrics can be greatly reduced using a dual-port memory, in which a memory bank for reading and writing a state metric of a first half among the N state metrics generated in a ACS unit and two memory banks for alternately reading and writing the state metric of the second half whenever a codeword is input.[156]

In a simplified trellis decoder a path metric value associated with each decoder state is updated upon the receipt of each incoming signal value. The path metric value is formed by:

1) Identifying those permissible transition(s) to that decoder state that are represented by symbols having the minimum branch metric compared to the received signal value
2) Identifying those permissible transition(s) to that decoder state that originate from states with the minimum previously-computed path metric value
3) Comparing the transitions identified in the first and second steps

From this comparison, the identity of the transition with the lowest path metric value is derived. Thereafter, the path metric chosen is stored for that given state.[157]

A reduced-state Viterbi detector employs a complement states grouping technique that comprises the steps of finding the state distances between complement states, forming a reduced-state trellis by grouping the complement states with state distance no less than the minimum free distance, and keeping the complement states with state distance less than minimum free distance unchanged. Generally, the technique causes no extra error propagation, which is a common problem for other reduced-state techniques. The complexity of the Viterbi detector is halved

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151 H. J. Nam and H. S. Kwak: "Viterbi Decoder for a High Definition Television", U. S. patent No. 5 844 945, 1 December 1998.

152 T. Hirose and H. Ino: "Viterbi Decoding Method and Viterbi Decoder" U. S. patent No. 5 946 329, 31 August 1999.

153 I. Hatakeyama: "Viterbi Decoder" U. S. patent No. 5 923 713, 13 July 1999.

154 I. Hatakeyama, "Viterbi Decoder" U. S. patent No. 5 982 822, 9 November 1999.

155 S. Ono and H. Katsuragawa: "Viterbi Decoding Method and Apparatus with Balance Among Memory and Processing Requirements", U. S. patent No. 5 787 127, 28 July 1998.

156 S. H. Choi and J. J. Kong, "Viterbi Decoder", U. S. patent No. 6 317 472, 13 November 2001.

157 V. Parizhsky: "Simplified Trellis Decoder", U. S. patent No. 5 588 028, 24 December 1996.

compared to the full-state Viterbi detector, with negligible loss in performance.[158] There has been design effort to reduce the number of times that the ACS processor must access memory, thereby to reduce the power required in the Viterbi decoding operations.[159]

The Viterbi decoder can be constructed so as to provide symbol synchronization.[160]

Trellis decoder architectures have been developed that permit seamless switching between different types of received digital modulation.[161]

6.16.1 Bibliography of Documents Relating to Trellis Decoders not Specifically Cited

D. I. Oh, M. S. Kim, and W. J. Lee: "Trellis Decoder of a DTV", U. S. patent No. 5 991 343, 23 November 1999.

R. W. Citta and D. A. Wilming: "Receiver for a Trellis Coded Digital Television Signal", U. S. patent No. 5 636 251, 3 June 1997.

Y. Okamoto: "Viterbi Decoder", U. S. patent No. 6 304 617, 16 October 2001

Y. Sasagawa: "Viterbi Decoder", U. S. patent No. 6 324 226, 27 November 2001.

6.16.1.1 Additional Bibliography Regarding Memory Organization in Trellis Decoding

F. C. Chen and R. G. Batruni: "Viterbi Decoder Circuit", U. S. patent No. 5 828 675, 27 October 1998.

6.16.1.2 Additional Bibliography Regarding ACS Circuitry in Trellis Decoding

C. Y. Tsui and R. S. K. Cheng: "Circuit for use in a Viterbi Decoder", U. S. patent No. 6 070 263, 30 May 2000.

D. Hansquine: "High-speed ACS for Viterbi decoder implementations", U. S. patent No. 6 333 954, 25 December 2001.

J. Meyer: "Convolution Decoder Using the Viterbi Algorithm", U. S. patent No. 5 802 115, 1 September 1998.

Lee and J. L. Sonntag: "Add Compare Select Circuit and Method Implementing a Viterbi Algorithm", U. S. patent No. 6 148 431, 14 November 2000.

M. Cartier: "Signal Processing Circuit to Implement a Viterbi Algorithm", U. S. patent No. 5 881 106, 9 March 1999.

M. Zarubinsky, Y. Salant, and N. Baron: "Apparatus and Method for Implementing Viterbi Butterflies", U. S. patent No. 6 257 756, 10 July 2001.

R. Ramesh and G. Kellam: "Non-binary viterbi decoder using butterfly operations", U. S. patent No. 6 115 436, 5 September 2000.

T. Andoh: "Viterbi Decoder with Pipelined ACS Circuits", U. S. patent No. 6 259 749, 10 July 2001.

T. Ishikawa and H. Suzuki: "Arithmetic Apparatus for use in Viterbi Decoding", U. S. patent No. 5 970 097, 19 October 1999.

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158 R. He and J. R. Cruz: "Implementing Reduced-state Viterbi Detectors", U. S. patent No. 6 081 562, 27 June 2000.

159 Y. Sasagawa: "Viterbi Decoder", U. S. patent No. 6 324 226, 27 November 2001.

160 T. Ikeda, Y. Ikeda, and T. Okada: "Viterbi Decoding Method and Decoder Capable of Eliminating Phase Indeterminacy", U. S. patent No. 5 724 394, 3 March 1998.

161 K. Hu, W. W. L. Lin, and M. D. Caldwell: "Digital Packet Data Trellis Decoder", U. S. patent No. 5 914 988, 22 June 1999.

Y. B. Choi: "Add-compare-select processor in a Viterbi Decoder", U. S. patent No. 5 928 378, 27 July 1999.

Y. Saegusa: "Viterbi Decoder", U. S. patent No. 6 343 105, 29 January 2002.

6.16.1.3 Additional Bibliography Regarding Branch Metric Computation and Storage

J. Murayama: "Viterbi Decoding Apparatus and Viterbi Decoding Method", U. S. patent No. 6 301 314, 9 October 2001.

J. S. Snyder, Jr.: "Apparatus and Method for Calculating Viterbi Path Metric Using Exponentially-Weighted Moving Average", U. S. patent No. 5 841 796, 24 November 1998.

J. Seflur and H. Tran: "Efficient Metric Memory Configuration for a Viterbi Decoder", U. S. patent No. 6 438 181, 20 August 2002.

K. Fujimoto: "Viterbi Decoder and Viterbi Decoding Method", U. S. patent No. 6 148 431, 14 November 2000.

R. A. Cesari and S. Simanapalli: "Method of efficient branch metric computation for a Viterbi convolutional decoder", U. S. patent No. 5 912 908, 15 June 1999.

R. A. Cesari: "Viterbi Decoder with Reduced Metric Computation", U. S. patent No. 5 844 947, 1 December 1998.

6.16.1.4 Further Bibliography Regarding Traceback and Survivor Selection

D. Dabiri, D. A. Luthi, and A. M. Mogre: "Area-efficient Surviving Paths Unit for Viterbi Decoders", U. S. patent No. 5 996 112, 30 November 1999.

D. I. Oh, E. R. Kim, D. H. Kim, and W. J. Lee: "Trellis Decoder for ATSC 8-VSB", U. S. patent No. 6 031 876, 29 February 2000.

D. I. Oh: "Traceback Device of a Trellis Decoder", U. S. patent No. 6 075 822, 13 June 2000.

D. M. Blaker, G. S. Ellard, and M. S. Mobin: "Decreasing Length Tracebacks", U. S. patent No. 5 533 065, 2 July 1996.

G. Feygin, et al.: "Architectural Tradeoffs for Survivor Sequence Memory Management in Viterbi Decoders", *IEEE Trans. on Communications*, vol. 41, no. 3, March 1993.

H. Jekal: "Traceback Processor for Use in a Trellis-coded Modulation Decoder", U. S. patent No. 6 134 697, 17 October 2000.

M. J. Rim and I. Oh: "Traceback-Performing Apparatus in Viterbi Decoder", U. S. patent No. 5 712 880, 27 January 1998.

N. Sayiner and J. L. Sonntag: "Viterbi Detector Using Path Memory Controlled by Best State Information", U. S. patent No. 6 097 769, 1 August 2000.

S. Araki, Y. Shimazaki, and S. Ono: "Viterbi Decoding Method and Circuit with Accelerated Back-Tracing and Efficient Path Metric Calculation", U. S. patent No. 5 946 361, 31 August 1999.

S. Chennakeshu, R. D. Koilpillai, and J. B. Anderson: "Method and Apparatus for Symbol Decoding", U. S. patent No. 5 905 742, 18 May 1999.

Y. B. Choi: "Trace-back Method and Apparatus for Use in a Viterbi Decoder", U. S. patent No. 5 878 092, 2 March 1999.

6.16.2 Partial-Response Trellis Decoding

A partial-response trellis decoder is used when a comb filter for suppressing NTSC co-channel interference is inserted into the path of the baseband DTV signal to perform partial-response pre-coding and the partial-response post-coding is deferred until after trellis decoding.[162] [163] Such deferral reduces the chance of error in the partial-response post-coding, but decreases the "eye" opening to noise ratio.

Viterbi trellis decoders that can seamlessly decode either 8-state or 4-state trellis coding are known.[164] Such trellis decoding is suitable when comb filtering to reject NTSC co-channel interference is performed on baseband DTV signal. Viterbi trellis decoders are known that can decode a trellis-encoded stream, using either an 8-state maximum likelihood response decoding mode for signal received via a Gaussian channel or a 16-state partial response decoding mode for signal received via a partial response channel.[165] Such trellis decoding is suitable when comb filtering to reject NTSC co-channel interference is performed in the IF passband.

A look-ahead maximum likelihood (ML) detector optimized in light of the noise correlation generated by the partial response channel is alleged to provide performance comparable to, or better than, a Viterbi detector in the presence of colored noise.[166]

6.16.2.1 Bibliography of Documents Regarding Partial-Response Trellis Decoding not Specifically Cited

G. Fettweis, R. Karabed, P. H. Siegel, and H. K. Thapar: "Reduced-Complexity Viterbi Detector Architectures for Partial Response Signaling", pp. 559–563.

K. J. Hole and O. Ytrehus: "Improved Coding Techniques for Precoded Partial-Response Channels," *IEEE Trans. on Information Theory*, vol. 40, no. 2, pp. 482–493, March 1994.

R. Haeb: "A Modified Trellis Coding Technique for Partial Response Channels," *IEEE Trans. on Communications*, vol. 40, no. 3, pp. 513–520, March 1992.

6.16.3 Soft-Decision Trellis Decoder

A soft-decision decoder can be constructed that includes an adaptive equalizer and a reliability information generator. The reliability information generator, receiving soft-decision data and equalized square error from the adaptive filter, performs delay detection of the soft-decision data, and then rotates the phase of the delay detected data so as to match the data to transmitted data components. The rotated data is weighted with the average value of the equalized square error to generate the reliability information for the soft-decision Viterbi decoding. This makes it possible to obviate the exponential or logarithm calculations which are conventionally needed for generating the reliability information for carrying out the soft-decision Viterbi decoding of the output of the adaptive equalizer, and hence to reduce the enormous amount of calculations and the size of hardware like DSP for performing the calculations.[167]

The decoding of convolutional codes with soft-input and soft-output values is often performed according to the principle of the symbol-by-symbol MAP algorithm (MAP = Maximum A posteriori Probability). This decoding algorithm can be realized by the trellis diagram of the convolutional code when a forward and backward recursion is used. Both the forward recursion and the backward recursion are very similar to the Viterbi algorithm except for

[162] R. W. Citta and D. A. Wilming: "Trellis Coded Modulation System for HDTV", U. S. patent No. 5 583 889, 10 December 1996.

[163] D. J. Kim and H. S. Kwak: "Partial Response Trellis Decoder for High Definition Television (HDTV) System", U. S. patent No. 5 508 752, 16 April 1996.

[164] K. Hu, W. W. L. Lin, and M. D. Caldwell: "Viterbi Decoder for Digital Packet Signals", U. S. patent No. 5 841 819, 24 November 1998.

[165] H. Jekal: "Method and Apparatus for Decoding Trellis Code Data", U. S. patent No. 6 088 404, 11 July 2000.

[166] R. G. Yamasaki, R. Kuki, H. M. Lin, and C. Tammel: "Method and Apparatus for Reducing Noise Correlation in a Partial Response Channel", U. S. patent No. 5 995 561, 30 November 1999.

[167] H. Igarashi, "Soft Decision Decoder", U. S. patent No. 5 923 713, 13 July 1999.

the recursion direction. The accumulated metrics calculated during the backward recursion are to be stored, because they are necessary in the forward recursion for calculating the soft-output values. The memory requirement for this is $N(2^L-1)$ words (of 16 bits), where N is the block length and L the influence length of the convolutional code. Even with moderate block lengths N of several hundred bits, this implies a large memory requirement which cannot be satisfied in currently available digital signal processors. The exact symbol-by-symbol MAP algorithm is basically unsuitable for fixed point DSP's, because the algorithm needs as soft-input values probabilities whose combinations in the algorithm (multiplication and addition) will rapidly lead to numerical problems. A variant of the symbol-by-symbol MAP algorithm stores only the backward state metric in each L^{th} step, thereby reducing the memory requirement for the backward metric by a factor of L, where L is the influence length of the convolutional code. The results are transferred to a sub-optimum algorithm that utilizes log-likelihood ratios for further saving of memory and computational effort when soft-output values are only needed for selected bits. The soft-output algorithm is restricted to the number of soft-output bits, and the conventional Viterbi algorithm is used for the remaining bits.[168]

6.16.3.1 Bibliography of Document Regarding Soft-Decision Trellis Decoding but not Specifically Cited

F. Koizumi: "Maximum-likelihood Decoding", U. S. patent No. 5 995 562, 30 November 1999.

J. Hagenauer and P. Höher: "Method for Generalizing the Viterbi Algorithm and Devices for Executing the Method", U. S. patent No. 5 181 209, 19 January 1993.

J. Hagenauer, et al.: "A Viterbi Algorithm with Soft-Decision Outputs and Its Applications," Communications Technology for the 1990's and Beyond, Dallas, Institute of Electrical and Electronics Engineers, vol. 3, pp. 1680–1686, November 27–30, 1989.

S. Ono, H. Hayashi, T. Tanaka, and N. Kondoh: "Soft Decision Estimation Unit and Maximum-Likelihood Sequence Estimation Unit", U. S. patent No. 6 302 576, 16 October 2001.

6.17 Data De-Interleaver

The data de-interleaver is readily implemented in dual-port random-access memory.[169]

6.18 Reed-Solomon Decoder

6.18.1 R-S Correction of up to 20-Byte-Errors/Segment

The (207, 187) Reed-Solomon forward-error-correction code supports correction of ten byte errors in 207 bytes if the locations of bytes in error must be determined from the code itself. If the locations of bytes in error are known from other sources, the (207, 187) R-S FEC supports correction of up to twenty byte errors in 207 bytes, as much as doubling the maximum length of isolated continuous noise burst that can be corrected.[170] Soft-decision trellis decoding can supply the locations of bytes in error to the Reed-Solomon decoder.[171] [172] If the AGC of the receiver is

[168] J. Petersen: "Soft-Output Decoding Transmission System with Reduced Memory Requirement", U. S. patent No. 6 028 899, 22 February 2000.

[169] M. Fimoff, S. F. Halozan, and R. C. Hauge: "Convolutional Interleaver and Deinterleaver", U.S. patent No. 5 572 532, 5 November 1996.

[170] D. A. Luthi: "Video Device with Reed-Solomon Erasure Decoder and Method Thereof", U. S. patent No. 5 875 199, 23 February 1999.

[171] K. Miya: "Error Detection Method Using Convolutional Code and Viterbi Decoding", U. S. patent No. 5 530 708, 25 June 1996.

[172] H. Tanaka: "Reliability of Maximum-Likelihood Decoded Codewords and Its Application to Decoding of Concatenated Codes," Electronics and Communications in Japan, Part 3, vol. 76, no.4, pp. 71–78, 1993.

insensitive to impulse noise, impulse noise peaks can be detected to help determine the locations of bytes in error.

6.19 Transport Stream De-Multiplexing

Transport stream de-multiplexing apparatus sorts data packets to appropriate “back-end” decoders responsive to packet identification (PID) portions of their headers. The PID match filter for the respective packets designated to be selected to each decoder is either hard-wired or is programmed responsive to one or more of the PSI (Program Specific Information) tables temporarily stored in receiver memory. These PSI tables are periodically transmitted in data packets contained within the transport stream and provided with their own PIDs. These PSI tables are selected from the transport stream for being written into receiver memory.

6.19.1 Treatment of Data Segments Containing Uncorrected Error(s)

In receivers designed for the consumer market from 1995 until 2003, the conventional practice has been to discard data packets that are known to contain uncorrected error, as part of the transport stream de-multiplexing procedure. The MPEG-2 and AC-3 decoders used in these designs lacked the processing power to deal with data packets that are known to contain uncorrected error. An MPEG-2 or AC-3 decoder can detect the fact that data packets have been discarded by detecting discontinuities in the modular count of the type of data packets forwarded to that decoder, which modular count is contained in the headers of those data packets. The MPEG-2 or AC-3 decoder is usually designed to institute measures for concealing errors introduced by a data packet having been discarded, responsive to the decoder detecting that a data packet has been discarded.

In many receiver designs, Reed-Solomon error-correction circuitry capable of locating as many as 20 errors, but correcting only 10 at most, is arranged to supply indication of when its error-correcting capability is exceeded. This indication is used to identify data packets that contain uncorrected error.

In receiver designs in which the Reed-Solomon error-correction circuitry is capable of correcting as many as 20 errors, errors are located by analyzing the confidence factor for each final decision that the trellis decoder makes when estimating the transmitted value of a received symbol. If the number of low-confidence decisions sought to be corrected by the Reed-Solomon error-correction algorithm exceeds 20, the data packet is determined to contain non-correctable error.

The use of portions of data packets other than those known to contain uncorrected error requires considerable processing in an MPEG-2 decoder, tending to raise its cost too high for the consumer market. The use of portions of data packets other than those known to contain uncorrected error is more feasible in an AC-3 decoder or in a decoder for constructing tables in receiver memory. If data packets known to contain uncorrected error are transmitted to back-end decoding apparatus, there will be no discontinuities in the modular count of the type of data packets forwarded to that apparatus for signaling the presence of uncorrected error. So, the transport error indicator bit in the header of each of the data packets containing uncorrected error is modified for signaling the back-end decoding apparatus that there is uncorrected error in that particular data packet.[173] Additionally, the bytes generated by the transport stream de-multiplexing operations can be extended to nine bits, appending a bit indicative of whether or not each byte is to be considered to contain error. Such measures facilitate byte errors being

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[173] K. F. Pang and D. D. Neuman: “Packet Error Propagation for MPEG Transport Demultiplexers”, U. S. patent No. 5 579 317, 26 November 1996.

accurately located, even though variable delay is introduced by buffers for the back-end decoding apparatus.

In such a scheme, each byte can be conditionally determined to contain error if trellis decoding generates a low confidence factor concerning any decision as to data contained in that byte. This conditional determination of error is reversed, if the Reed-Solomon error-correction circuitry can correct the byte or verify its being correct. Otherwise, the conditional determination of error is confirmed and becomes a final determination that the byte contains error.

6.20 DTV Receivers for both Over-the-Air VSB DTV Transmissions and other DTV Transmissions

DTV receivers for receiving QAM DTV transmissions over cable as well as VSB DTV transmissions over the air normally employ the same front-end up through the VHF IF amplifier for both types of signal. Different circuitry is used for demodulating the QAM DTV signal than is used for demodulating the VSB DTV signal, although the two types of demodulation circuitry often share common elements.[174] [175] [176] [177] [178] The equalization filtering for the different types of DTV signal may share elements in common.[179] [180] [181]

6.20.1 Detection of VSB Signal Reception

The reception of a VSB DTV signal rather than a QAM DTV signal can be determined by analyzing the average distribution of power in the power spectrum of the received signal to ascertain whether or not there is increased power density where the pilot carrier of the VSB signal would be located. Such measurements have been made using a sliding bi-quad filter.[182]

The detection of significant 5.38 MHz energy in the envelope of the received signal is an indication of VSB signal reception. Such detection can use narrowband filtering also used in a bright-spectral-line method of regenerating the symbol clock.

Another method for detecting VSB reception is to check the envelope of the received signal for the presence of PN511 sequence at 10.76 MHz symbol rate.[183] The received signal is envelope detected, and the envelope detection result is supplied to a match filter for the PN511 sequence included in the DFS signal. Detecting the presence of field-rate high-energy pulses in the match filter response is indicative of VSB reception.

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174 R. L. Cupo: “Sampling System for Radio Frequency Receiver”, U. S. patent No. 5 841 814, 24 November 1998.

175 C. B. Patel and A. L. R. Limberg: “Radio Receiver for Receiving Both VSB and QAM Digital HDTV Signals”, U. S. patent No. 6 104 442, 15 August 2000.

176 C. B. Patel and A. L. R. Limberg: “Radio Receivers For Receiving Both VSB and QAM Digital Television Signals with Carriers Offset by 2.69 MHz”, U. S. patent No. 6 333 767, 25 December 2001.

177 A. L. R. Limberg: “Television Receiver with Separate I-F Amplifiers for VSB and QAM Digital TV Signals That Are Digitally Synchrodyned”, U. S. patent No. 6 351 290, 26 February 2002.

178 A. L. R. Limberg: “TV Receiver Using Read-Only Memory Shared During VSB and QAM Reception for Synchrodyning I-F Signal to Baseband”, U. S. patent No. 6 496 229, 17 December 2002.

179 Y. Juan: “FIR Filters with Multiplexed Inputs Suitable for Use in Reconfigurable Adaptive Equalizers”, U. S. patent No. 5 642 382, 24 June 1997.

180 A. L. R. Limberg and C. B. Patel: “TV Reception Apparatus Using Same Ghost-Cancellation Circuitry for Receiving Different Types of TV Signals”, U. S. patent No. 6 313 882, 6 November 2001.

181 C. B. Patel and A. L. R. Limberg: “DTV Receiver with Baseband Equalization Filters for QAM Signal and for VSB Signal Which Employ Common Elements”, U. S. patent No. 6 313 885, 6 November 2001.

182 C. G. Scarpa: “Automatic VSB/QAM Modulation Recognition Method and Apparatus”, U. S. patent No. 5 557 337, 3 June 1997.

183 A. L. R. Limberg: “Envelope Detection of PN Sequences Accompanying VSB Signal to Control Operation of QAM/VSB DTV Receiver”, U. S. patent No. 6 480 236, 12 November 2002.

6.20.2 Detection of QAM or QPSK Signal Reception

The detection of significant 2.69 MHz energy in the envelope of the received signal, but very little 5.38 MHz energy, is an indication of QAM or QPSK signal reception. There are techniques for determining the type of QAM signal being received based on the probability density function of the received signal over a predetermined time period.[184] The received signal is squared and normalized to a preset value. Then, a histogram of the normalized squared signal is prepared. The constellation size of the QAM signal is discernible from the histogram. The histogram of a VSB DTV signal with accompanying pilot can also be distinguished from the histograms of QAM and QPSK signals.

6.20.3 Discrimination Between Various forms of Signal Using Midband Carrier

Comparison of the decision-feedback errors for different forms of hard-decision data slicing provides for discriminating which of various forms of signal using midband carrier is currently being received.[185] [186] [187]

6.21 DTV Receivers with NTSC Signal Reception Capability

TV receivers that can receive NTSC analog DTV signals as well as VSB DTV signals are apt to be constructed until there is no longer any transmission of NTSC analog DTV signals. These receivers are likely to use the same RF amplifier and first converter for both NTSC analog DTV signals and VSB DTV signals. Optimal IF amplifier designs for the two types of signal differ in regard to passband characteristics and to AGC. So, until such time as there is negligible over-the-air broadcasting of NTSC analog DTV signals, receivers may use different IF amplifiers for analog and digital TV signals.[188] [189] TV receivers have been proposed that can receive NTSC analog TV signals as well as VSB DTV signals that use the same IF amplifiers for both types of signals and that detect video modulation using elements also used in DTV demodulation.[190]

If there is a de-ghoster for the NTSC analog DTV signals, it may use elements in common with the equalizer for the VSB DTV signals.[189]

Frequency-synthesis techniques are the preferred way to generate beat-frequency oscillations for use in the first converter. When the channel is initially tuned, the frequency synthesizer can generate beat-frequency oscillations for converting a DTV signal to prescribed intermediate frequencies for DTV reception. Then, if DTV signal fails to be detected, the frequency synthesizer can generate beat-frequency oscillations for converting an NTSC signal to prescribed intermediate frequencies for NTSC reception.[188] Alternatively, when the channel is initially tuned, the frequency synthesizer can generate beat-frequency oscillations for converting an

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[184] F. A. Lane: “Technique For Automatically Detecting The Constellation Size Of A Quadrature Amplitude Modulated (QAM) Signal”, U. S. patent No. 5 557 337, 10 January 1995.

[185] J. S. Stewart: “Apparatus for Decoding Video Signals Encoded in Different Formats”, U. S. patent No. 5 666 170, 9 September 1997.

[186] J. S. Stewart: “Apparatus for Demodulating and Decoding Video Signals Encoded in Different Formats”, U. S. patent No. 5 671 253, 23 September 1997.

[187] J. S. Stewart: “Apparatus for Demodulating and Decoding Satellite, Terrestrial and Cable Transmitted Digital Television Data”, U. S. patent No. 5 717 471, 10 February 1998.

[188] R. W. Citta: “Dual HDTV/NTSC Receiver Using Sequentially Synthesized HDTV and NTSC Co-channel Carrier Frequencies”, U. S. patent No. 5 283 653, 1 February 1994.

[189] A. L. R. Limberg: “Radio Receiver Detecting Digital and Analog Television Radio-Frequency Signals with Single First Detector”, U. S. patent No. 5 982 457, 9 November 1999.

[190] C. H. Strolle and S. T. Jaffe: “Multiple Modulation Format Television Signal Receiver System”, U. S. patent No. 6 005 640, 21 December 1999.

NTSC signal to prescribed intermediate frequencies for NTSC reception. Then, if NTSC signal fails to be detected, the frequency synthesizer can generate beat-frequency oscillations for converting a DTV signal to prescribed intermediate frequencies for DTV reception.

Channel-addressed programmable memory can store indications of which channels convey NTSC broadcasts and which convey VSB DTV broadcasts, which indications are used for selectively activating portions of the receiver.[188].[191]

6.21.1 Detection of NTSC Signal Reception

The reception of an NTSC signal can be detected relying on the power spectrum of a VSB DTV signal supposedly being quite uniform across the middle five MHz or so of the six MHz transmission channel.[192] A first power measurement is made in a sub-band where the NTSC power spectrum normally has significant spectral energy, and a second power measurement is made in a sub-band where the NTSC power spectrum normally has relatively low spectral energy. If the first power measurement exceeds the second power measurement by a large enough factor, the received signal is determined to be an NTSC signal. Otherwise, the received signal is determined to be a DTV signal. Variants of the method which compare the power in more than two sub-bands offer increased reliability when multipath distortion affects the power spectrum.

Detection of the presence of 4.5 MHz intercarrier in the received signal is a trustworthy indication of the reception of NTSC analog TV signal.[193] Detection of VSB pilot carrier in the received signal as well is indicative of co-channel interference of NTSC analog TV and VSB DTV signals. The energies of the 4.5 MHz intercarrier and the VSB pilot signal can be compared for determining which of the NTSC analog TV and VSB DTV signals is the stronger.

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[191] D. J. Duffield and R. D. Altmanshofer: “Television Receiver having the Capability to Associated and HDTV and an NTSC Channel”, U. S. patent No. 5 461 427, 24 October 1995.

[192] C. G. Scarpa: “Automatic Television Signal Detector to Differentiate NTSC Signals from HDTV/ATV Signals”, U. S. patent No. 5 557 337, 16 September 1996.

[193] A. L. R. Limberg: “NTSC/DTV Reception Apparatus Automatically Conditioned for NTSC SIGNAL Reception Responsive to 4.5 MHz Intercarrier”, U. S. patent No. 6 307 595, 23 October 2001.

Annex A:
Guide to Existing TV Receiver Requirements and Standards

The following is a listing of mandatory and voluntary standards, recommended practices, and other reference information for conventional television receivers. A study of these materials will be helpful in understanding what changes will be necessary to accommodate DTV receivers and what may have to be added to cover DTV receivers, if necessary.

1. MANDATORY REQUIREMENTS

47, CFR, (FCC) Part 2, Marketing and Importation.

47, CFR, (FCC) Part 15, Radio Frequency Devices.

FCC/OET MP-2, Measurement of UHF Noise Figure of TV Receivers.

21CFR (FDA), Subchapter J, Radiological Health.

Part 1000, General.

Part 1002, Records and Reports.

Part 1020, Performance Standard for Ionizing Radiation Emitting Products.

Part 1020.10, Television Receivers.

ANSI C63.4, Standard Methods of Measurement of Radio-Noise Emissions From Low-Voltage Electrical And Electronic Equipment in The Range Of 9 kHz To 40 GHz.

FTC-16CFR, Part 410, Deceptive Advertising as to Sizes Of Viewable Pictures Shown By Television Receiving Sets.

1.1 Mandatory for "Cable Ready" Receivers

EIA IS-6, Recommended Cable Television Identification Plan.

EIA IS-23, RF Interface Specification for Television Receiving Devices and Cable Television Systems.

EIA-542-A, Channelization Plan for Cable Television Tuners.

1.2 Mandatory in Some States of the U.S. and/or in Canada

UL-1492, *Audio-Video Products and Accessories*.

UL-1413, *High Voltage Components for Television Type Appliances*.

UL-1418, *Implosion Protected Cathode Ray Tubes for Television Type Appliances*.

NFPA-70, *National Electrical Code*.

CSA-C22.2 No. 1-M90, *Radio, Television and Electronic Apparatus*.

2. VOLUNTARY STANDARDS

EIA IS-16A, Immunity of Television Receivers and Video Cassette Recorders (VCRs) to Direct Radiation From Radio Transmissions, 0.5 to 30 MHz.

EIA IS-31, Recommended Design Guideline - Rejection of Educational Interference to Ch. 6 Television Reception.

EIA-544-A, Immunity of TV and VCR Tuners to Internally Generated Harmonic Interference From Signals.

3. OTHER RELATED STANDARDS AND REFERENCES

3.1 FCC

47, CFR, (FCC) Part 73, Radio Broadcast Services.

47, CFR, (FCC) Part 76, Cable Television Service.

3.2 Safety AND x-Ray

NOM-001, Electronic Apparatus-Household Electronic-Apparatus by Different Sources of Electrical Power-Safety Requirements and Testing-Methods for Type Approval.

CPEB1, Standard Method of Measurement of Ionizing Radiation from Television Receivers for Factory Quality Assurance.

CPEB2, Definition of Normal Operating Conditions for Television Receivers.

CPEB3, Measurement Instrumentation for X-Radiation from Television Receivers.

EIA-500-A, Recommended Practice for Measurement of X-Radiation from Projection Cathode Ray Tubes.

EIA-503-A, Recommended Practice for the Measurement of X-Radiation From Direct View Television Picture Tubes.

3.3 Interference and Immunity

EIA-378, Measurement of Spurious Radiation From FM and TV Broadcast Receivers in The Frequency Range of 100 To 1000 MHz - Using The EIA Laurel Broadband Antenna.

CISPR 13, Limits and Methods of Measurement of Radio Interference Characteristics of Sound and Television Broadcast Receivers and Associated Equipment.

CISPR 20, Limits and Methods of Measurement of Immunity Characteristics of Sound and Television Broadcast Receivers and Associated Equipment.

IEEE-187-90, IEEE Standard on Radio Receivers: Open Field Method of Measurement of Spurious Radiation from FM and Television Broadcast Receivers.

3.4 Cathode Ray TV Display Tubes

EIA-256-A, Deflecting Yokes for Cathode Ray Tubes.

EIA-266-A, Registered Screen Dimensions for Monochrome Picture Tubes.

EIA-324-A, Registered Screen Dimensions for Color Picture Tubes.

EIA-493, Recommended Practice for Conversion of U. S. to Metric Dimensions for Color and Monochrome Cathode Ray Tubes and their Component Parts.

EIA-527, Screen Definition for Color Picture Tubes.

3.5 Voluntary TV Receiver Recommended Practices

EIA-462, Electrical Performance Standards for Television Broadcast Demodulators.

EIA-563, Standard Baseband (Audio/Video) Interface Between NTSC Television Receiving Devices And Peripheral Devices.

REC-109-CH, Intermediate Frequencies for Entertainment Receivers.

TVSB1, EIA Recommended Practice for use of a Vertical Interval Reference (VIR) Signal.

TVSB3, A History of the Vertical Interval Color Reference Signal (VIR).

TVSB5, Multichannel TV Sound System - BTSC System Recommended Practices.

3.6 International Standards, Including IEC

IEC 65, Safety Requirements for Mains Operated Electronic And Related Apparatus for Household and Other Similar General Use.

IEC 107-1 to -6, Methods of Measurement on Receivers for Television Broadcast Transmissions, (6 Parts).

IEC 569, Informative Guide for Subjective Tests On Television Receivers.

IEC 68, Environmental Testing - Part 1: General and Guidance.