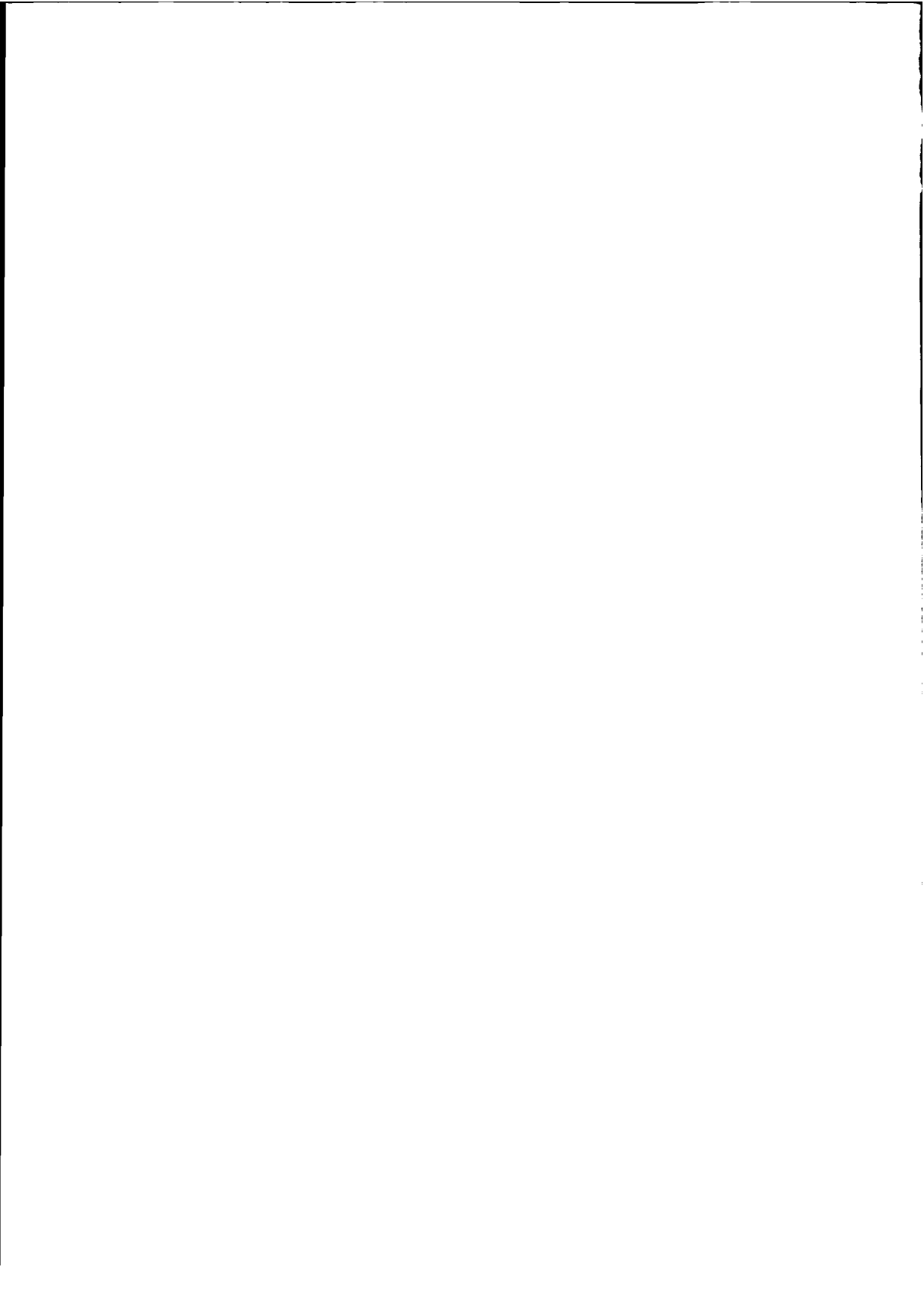


STELLINGEN

- 1c. The process of arranging a contest between competing proponents of new infrastructural systems through field and laboratory tests as often used in the U.S.A. is beneficial, provided that theoretical discussions about the technical merits of the proposals are banned from the agenda of committees that prepare and define the tests.
- 2c. During field and laboratory tests of competing infrastructural systems representatives of the proponent companies should have access to the test sites at all times.
- 3t. A Ghost Canceling Reference signal should be analog regardless of whether the information that it is intended to protect is analog or digital.
- 4t. Tests of a Ghost Cancellation Reference signal should be performed with professional system implementations of the highest possible quality in order to avoid as much as possible the possibility that the implementation rather than the GCR is tested.
- 5n. The fact that most people, in the U.S.A. at least, are unaware of the relation between national savings and trade balance causes fruitless discussions.
- 6n. The Cultural Revolution in China seems to have caused many fatal cases of cancer among leading people in that country.

- 7t. The research program on Echo Cancellation on which this thesis reports benefited to a considerable extent from the fact that a number of participants each were knowledgeable in analog electronics, digital electronics as well as computer science.
- 8c. It is amazing that even in the exact sciences wrong ideas are quoted and re-quoted from the literature in an uncritical fashion. Because of this it is desirable for individual researchers to derive or confirm their insights from basic theory by themselves.
- 9n. In many cases one can gain more insight from reading and rereading truly outstanding older literature than from trying to keep up with the far more abundant current literature.
- 10n. If reaching a decision is found to be difficult one should be aware of the possibility that the choice among the various alternatives is unimportant. This may save time and money.
- 11n. It is desirable that industrial researchers have a working knowledge of the principles of corporate finance. This would facilitate finding the best strategy to secure proper support from the right people at the right time.
- 12c. The diversity of the nations of origin of workers in many American industrial laboratories impedes acceptance of their results by public and in house committees because their ability to operate and compete effectively in the local arena often is inferior to their technical ability. Management should be keenly aware of this.

- 13n. Reacting in the working place to the "noise" generated by the abundant amount of up to date information that nowadays is available to everyone, everywhere and at any time in fact constitutes unnecessary and fruitless work that makes an irrelevant contribution to the Gross Domestic Product and thus falsifies the "productivity" of a nation.
- 14n. Provincial attitudes among the general population of the U.S.A. are inconsistent with the prominent international position of that country.



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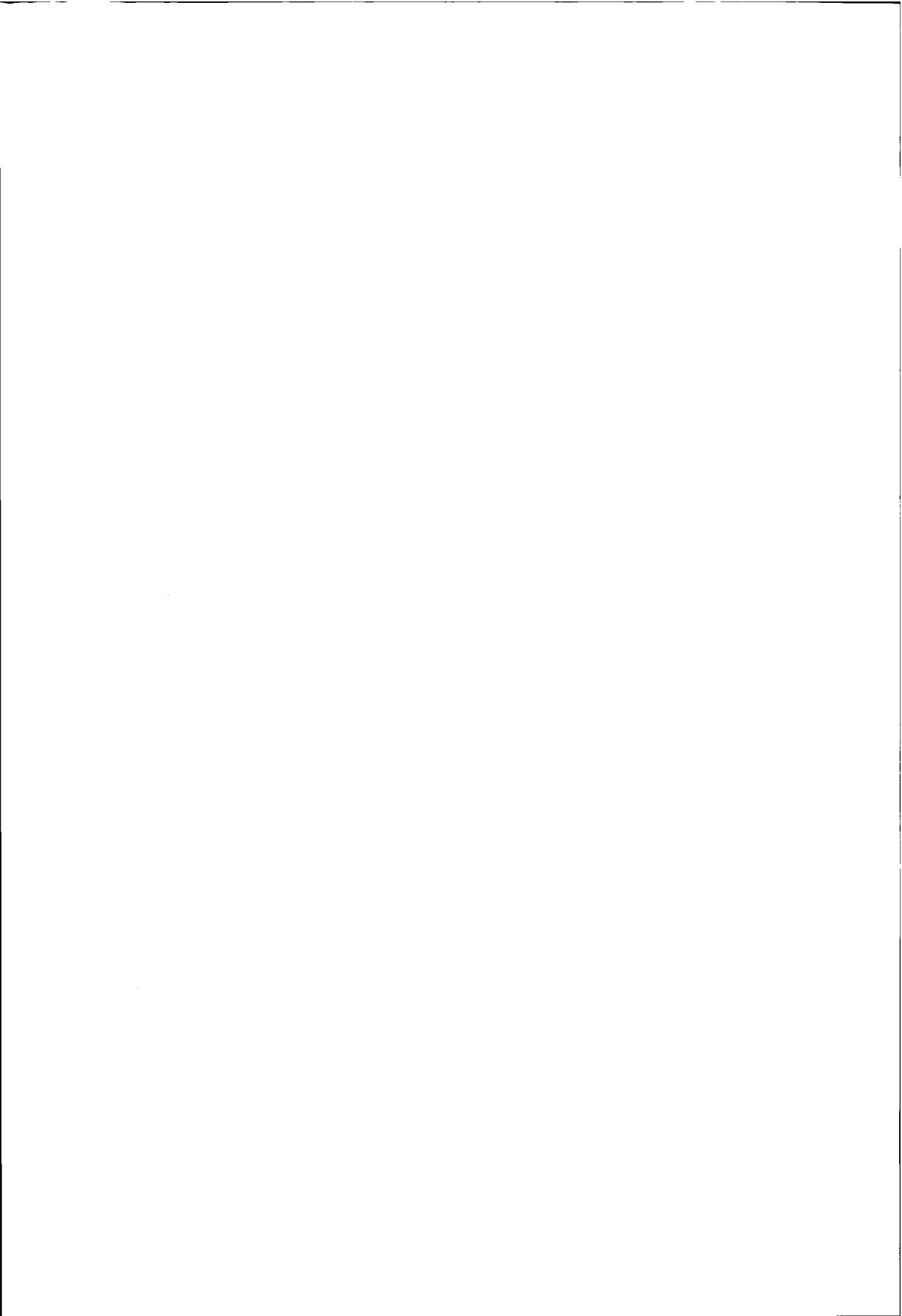
STELLINGEN

1. Het proces om middels testen in het veld en in laboratoria een wedstrijd te organiseren tussen de concurrerende organisaties die voorstellen hebben ingediend voor systemen van infrastructurele aard, zoals gebruikelijk in de Verenigde Staten, werkt gunstig mits theoretische discussies over de kwaliteiten van die voorstellen niet worden toegelaten op de agenda van commissies die de testen voorbereiden en definiëren.
2. Gedurende het testen van concurrerende infrastructurele systemen in het veld en in laboratoria moeten vertegenwoordigers van de voorstellende firma's te allen tijde toegang hebben tot de plaatsen waar de testen worden verricht.
3. Een echo-onderdrukkingsreferentiesignaal moet analoog zijn, onafhankelijk van de vraag of de informatie die beschermd moet worden analoog of digitaal is.
4. Testen van een echo-onderdrukkingsreferentiesignaal behoren te worden verricht met professionele systeemimplementaties van de hoogst mogelijke kwaliteit ten einde zoveel mogelijk te voorkomen dat de implementatie in plaats van het referentiesignaal wordt getest.
5. Het feit dat een meerderheid onder de bevolking, in de Verenigde State in ieder geval, zich niet bewust is van de relatie tussen nationale besparingen en de handelsbalans, veroorzaakt vruchteloze discussies.
6. De Culturele Revolutie in China heeft waarschijnlijk vele fatale gevallen van kanker veroorzaakt onder leidende functionarissen in dat land.
7. Het onderzoeksprogramma over Echo Onderdrukking, waarover dit proefschrift rapporteert, werd in belangrijke mate gediend door het feit dat een aantal medewerkers ieder voor zich vertrouwd waren met zowel analoge elektronica, digitale elektronica en computertechniek.
8. Het is verbazingwekkend dat zelfs in exacte wetenschappen onjuiste ideeën uit de literatuur op onkritische wijze worden aangehaald en wederom aangehaald. Het is daarom wenselijk dat individuele onderzoekers zelf hun inzichten afleiden of bevestigen op basis van fundamentele theorie.
9. In vele gevallen kan men meer inzicht verkrijgen middels het lezen en herlezen van uitstekende oude literatuur dan middels het bijhouden van de veel uitgebreidere recente literatuur.
10. Wanneer het bereiken van een beslissing moeilijk blijkt te zijn moet men zich rekenschap geven van de mogelijkheid dat de keuze tussen de verschillende opties van weinig belang is. Dit kan tijd en geld besparen.
11. Het is wenselijk dat onderzoekers die werkzaam zijn in de industrie enige notie hebben van de financiële inzichten en methodes die daar gebruikelijk zijn.
12. De diversiteit aan nationaliteiten van medewerkers van vele Amerikaanse laboratoria bemoeilijkt de acceptatie van hun resultaten door openbare en lokale commissies omdat hun vaardigheid om effectief in de lokale arena te functioneren en te wedijveren vaak inferieur is vergeleken met hun technische kundigheden. De leiding van die laboratoria moet zich hier zeer bewust van zijn.
13. Het op de werkplaats reageren op de "ruis" die wordt gegenereerd door de grote hoeveelheid recente informatie die tegenwoordig overal en voortdurend voor iedereen beschikbaar is, veroorzaakt onnodig en vruchteloos werk dat een irrelevante bijdrage levert tot het Nationaal Produkt en dus een falsificatie vormt van de "productiviteit" van een land.
14. De provinciale instelling die heerst onder brede lagen van de bevolking van de Verenigde Staten strookt niet met de prominente internationale positie van dat land.

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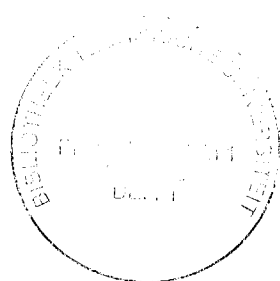
TR3147

Echo Cancellation in Television



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Echo Cancellation in Television



PROEFSCHRIFT

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in het openbaar te verdedigen ten overstaan van een commissie,
door het College voor Promoties aangewezen,
op dinsdag 9 juni 1998 te 10:30 uur

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David KOO

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TABLE OF CONTENTS

Abstract	3
Ch.1: Introduction	4
Ch.2: System Level Considerations	11
Ch.3: Preliminary Introduction to Echoes	37
Ch.4: Echoes in Television Channels	45
Ch.5: Requirements for Ghost Cancellation Reference signals (GCRs)	80
Ch.6: The Design of a Family of GCRs	92
Ch.7: Considerations on the Use of Digital Sequences as GCRs	101
Ch.8: Algorithms for Channel Characterization	115
Ch.9: Programmable Filters for Echo Cancellation	126
Ch.10: Overall Cancellation System and Simulation	135

Ch.11: Implementation of the Hardware	145
Ch.12: Implementation of the Software	153
Ch.13: Potential Reasons for Failure of the System	166
Ch.14: Tests and Conclusions on Competing Systems in the USA (1992) .	170
Ch.15: International Aspects	180
Ch.16: Prospects for Further Work	185
Ch.17: Conclusions	187
Appendix	189
Summary	202
Samenvatting (Summary in Dutch)	204
Acknowledgement	206
Biography	207

ABSTRACT

In this thesis the subject of echo cancellation in television is treated systematically, including the contributions that the author has been able to make. A thorough analysis of the creation of echoes, taking into account various modulation methods and their consequences as to the in-phase as well as the quadrature channel, leads to the insight that this process is linear from base-band input to base-band output. This opens the way towards a formulation of the requirements that a Ghost Cancellation Reference (GCR) signal has to meet, which is used for the design of an optimal family of GCR's. The adjustable parameters that are introduced in this design leave sufficient room for accommodation to the various television systems that are in use today. Verification by computer simulation is reported, to show that these GCR's indeed provide excellent cancellation.

Subsequently the actual design of hardware and software for a full cancellation system that was used in field tests for various organizations in a number of countries is described and a digest of the test results for this system and, in the case of the U.S.A., the competing systems is given. The test results led to the elevation of our GCR to U.S.A. standard by the Federal Communications Commission and to world standard by the International Telecommunications Union, as well as to the Emmy award that the author received for this work in 1995.

CHAPTER 1

INTRODUCTION

In the field of Consumer Electronics the impact of the advent of Integrated Circuits (the so-called "chips") has caused a major revolution, much like it did in the Computer field. In the first place the number of transistors and other circuit elements that a designer can use in a single TV set has increased so much that it is nearly inconceivable: when electronic tubes or, later, separate transistors were still used in radios and TV sets a number of 20 of these devices in a single TV set was considered large. Nowadays we use Integrated Circuits, or chips that contain millions of them, all concentrated on a piece of Silicon of perhaps 1 square centimeter. In the second place the opportunity to do digital signal processing has arisen, in which the various communication signals that we use, like TV or Radio signals, are first translated into a series of 0 and 1 symbols which can then be processed by computer-like devices. These devices differ from the old ones because, again like in a computer, they can be programmed and thus are flexible. We even can use a normal (micro) computer to calculate what the programming has to be. This can even be done "on the fly", that means "as fast as we need it", without being hampered by the slowness of a the human operator or programmer. In the third place the cost of all this to the consumer has been very low: the price of a TV set has not essentially changed while its functionality has increased manyfold.

One of the new possibilities is the removal of "ghosts" or "echoes" from TV pictures. These form a real problem amidst mountains and in cities with high-rise buildings, especially when a TV antenna is used for reception of the TV signal. The ghosts come about by reflection of the waves that carry the TV signal, against these mountains and buildings. The problem with their removal is that the manufacturer of the TV set cannot possibly predict what the ghosts will look like for each channel in each city in each building at all times in the future. He could therefore not design a really well working ghost canceling system into his receivers, before the advent of programmable Integrated Circuits.

Nowadays this can be done because the chips in the receiver can be designed and be programmed to be capable of looking at the ghosted signal by themselves and then to calculate how other integrated circuits should be programmed to remove the ghosts. This programming can even be done so quickly that in a few seconds after for instance a channel change the ghosts in the new channel have been canceled.

This thesis is about this particular new development, in which the author has been able to deliver contributions that were recognized by an Emmy Award in the United States and led to a world standard for the standard signal (of his invention) that has to be used in the process, and that the broadcasters have to send all the time between the TV pictures that come out of their cameras.

From the beginning of TV onwards there have been concerns about echoes, in particular amidst high rise buildings and mountains and it was realized early on that the use of some kind of standardized Ghost Canceling Reference signal could be a solution. The normal synchronization signals of TV were not sufficiently uniform for this apart from the fact that they were not designed for the purpose and

therefore not optimal. It has been suggested [1] that in countries where teletext is present in the vertical blanking interval (see chapter 2) the fact that in the absence of echoes this is a strictly two level signal can be used to formulate an algorithm for cancellation of echoes. At the time this was attempted with the aid of integrated gyrators but in fact it was too early. In 1989 a serious effort had come to fruition in Japan where extensive tests were done on competing systems and GCRs in order to arrive at a national standard [2]. Here the idea that a good GCR should have at least a smooth, if not white, spectrum got established for the first time to the best of our knowledge[2]. As a result of the standardization in Japan an effort in the USA was launched to establish a domestic standard. Various companies were allowed to present proposals to be tested in 1992 in a series of tests[3][4][5][6][7]. Eventually our system won these.

The aim of the work reported in this thesis has thus been to design an optimal "Ghost Cancellation Reference" (GCR) signal and to build a real ghost cancellation system, based on this GCR, for professional use in the broadcast industry, from which consumer versions can be derived.

The actual contributions that we made can be briefly summed up as follows:

1. A list of requirements for GCRs was formulated that was complete enough to allow for a well defined design of a family of GCRs with a number of adjustable parameters that was sufficient to attain adequate specialization for the various TV systems that are in use in the world.

2. It could be shown analytically and by computer simulation that Pseudo Random Digital Sequences do not meet these requirements.

3. Actual GCRs for NTSC, as used in the USA, and for several other TV systems were created.

4. A practical system, to be used for tests in the USA and various other countries was built. This system contains a substantial amount of software to ensure that the GCR is used in an optimal way, even under adverse conditions.

5. Considerable work was done to assist in and to obtain standardization of the GCR internationally.

As to the scope of the work, we could base ourselves on a theoretical study of the creation of ghosts in actual TV channels, taking into account various modulation systems (including Vestigial Side Band, or VSB modulation as commonly used in TV). This study showed that from base band to base band this is a linear process. This gave us the opportunity to restrict ourselves to linear correction methods with the aid of programmable digital filters and at the same time imposed on us the condition that we could only consider a linear TV chain, including linear, or synchronous detection in the receiver. Obviously FM modulated signals, as used in satellite TV, could not be considered.

Further restrictions on the scope of the work were imposed by trade off considerations in the actual construction of the system that were dictated by price/performance considerations on the length of filters, processing power, speed, accuracy etc.

The thesis starts with a general introduction which is this Chapter 1. The discussion on overall system considerations is in Chapter 2. A preliminary description of echoes in a natural channel of infinite bandwidth as well as in a real TV

channel is given in Chapter 3, whereas the creation of echoes in a real TV channel for various modulation systems of increasing complexity is in Chapter 4.

Chapter 5 is used to formulate the requirements on GCRs that have to be observed in their design. Chapter 6 contains the reasoning that led to the design of a single family of GCRs for the various TV systems in the world.

Chapter 7 constitutes a digression about Pseudo Random Digital Sequences, showing that these are not suitable as GCR.

A number of possible algorithms for Channel Characterization are described in Chapter 8. Chapter 9 is a treatment of actual Echo Cancellation, using programmable digital filters. Chapter 10 describes the overall structure of the system that we built as well as our simulation results on the expected operation of that system. A more detailed description of the hardware of the system is in Chapter 11. Chapter 12 gives an account of the software of the system.

A series of reasons for which the system may fail in optimally canceling echoes on certain occasions is discussed in Chapter 13. Chapter 14 reports on the USA tests and their results. It also includes some discussions on potential reasons for system failure and thus provides an explanation for the fact that the test results were not totally perfect although considerably better than those obtained by competing systems.

Chapter 15 is a brief account of the international standardization that occurred after national standardization in the USA was achieved. Chapter 16 is used to make a few remarks about prospects for the future. Chapter 17 contains the conclusions that may be drawn from the present work.

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CHAPTER 2

SYSTEM LEVEL CONSIDERATIONS

The discussions in this chapter serve two purposes. In the first place the chapter has to serve as a general introduction (section 2.1) to those aspects of current TV technology that are important for understanding the rest of this thesis. In the second place we give a summary treatment (sections 2.2 and 2.3) already in this early stage of issues that are important in echo cancellation and that can be discussed without demanding from the reader any detailed knowledge of the specific results of our work.

2.1 Short Description of Current Television Technology

Television is the broadcast of moving pictures and accompanying sound. The picture part of the TV signal is created in a TV camera that scans the image along horizontal lines for the three color components red, green and blue separately. From these three signals the system derives luminance and two other signals relating to color.

For luminance it has been determined in the past [1][2][3] that the human perceptual system requires a vertical resolution of about 500 lines for a picture viewed at a distance of about five times the picture height.

In practice some of these lines are transmitted without picture information in the time interval during which the scanning returns from the bottom of the picture to the top in preparation of the next scan. This time interval is commonly referred to as the "Vertical Blanking Interval" (VBI). Among the information transmitted during this VBI, an important element is the synchronization information [3] that tells the receiver when a new picture is started. Also, part of each line duration is used for synchronization information [3] to tell the receiver when a new line is started as well as for the so-called color burst [3] that informs the receiver of the phase of the color sub-carrier mentioned below.

In order to have the same horizontal resolution along the lines about 700 samples must be taken if the aspect ratio of the picture is 3:4. The human perceptual system also requires [1][3] a picture refresh rate of about 30 Hz. Thus the TV signal should be able to support a sampling rate of the order of $500 \times 700 \times 30$ which about equals 10.5 MHz, resulting in a bandwidth of the order of 5 MHz.

For color it has been established [3][4] that the required resolution is considerably less than for luminance. The two color signals mentioned above are modulated in quadrature [5][6] on a color subcarrier at (for NTSC-M) 3.58 MHz. For this a bandwidth of 2 MHz is interleaved inside the channel within the upper part of the frequency range that is also used for luminance [3][7][8]. In addition sound has to be transmitted for which a bandwidth of about 100 kHz is reserved [9][10] around a sound subcarrier. This subcarrier is located at the high frequency side of the TV channel, which thus has a total bandwidth of about 6 MHz.

The modulation method generally used for TV is Vestigial Side-Band modulation [11][12][13], to be explained in detail in Chapter 4, in which the major part of the information is sent in Single Side-Band and a minor part, including the dc component in Double Side-Band.

The resulting RF signal is reflected in an uncontrolled way against buildings, mountains etc. in terrestrial transmission over the air and also in cable systems, causing the echoes that are the subject of this thesis. Obviously these echoes will be different at different locations, antenna orientations and for different channels. Therefore echo cancellation has to be adaptive by nature. Nowadays that means that it has to be done by digital signal processing methods.

An important aspect of the cancellation process is that the receiver has to operate on the basis of information on what at least part of the signal has been before the distortion has occurred so that it "knows" the target of the restoration that has to be accomplished.

A historically interesting example has been tried in Europe [14] where a digital teletext [14] signal is transmitted in the VBI to provide textual information to viewers. This strictly two level signal becomes multi level in case of echoes and therefore the canceler can in principle be programmed to restore the two level character of the signal.

Another approach is to send a "Ghost Cancellation Reference Signal" (GCR) during the VBI which is also stored in undistorted form in the receiver. This method allows standardization, virtually independent of the presence or non-presence of other services and is therefore nowadays the preferred method. The receiver in this case has to restore the received GCR to be the same as the GCR in

its memory.

In this thesis we analyze the echo creation process, report on the design of a family of GCRs that became world standard by the actions of the International Telecommunications Union (ITU) and give an account of our computer verifications of its suitability as well as of the actual creation of an Echo Cancellation System that we used in the numerous field tests that have been conducted in various countries prior to and after the ITU decision. We also comment on some of the other GCRs that have been proposed.

2.2 Analysis

This section contains an analysis of various phenomena in echo cancellation that have to be taken into account even before one starts out on the design phase of an echo cancellation system.

2.2.1 Linearity

We will show in chapter 4 in great detail that even in the complicated but realistic case of Vestigial Side-Band modulation the creation of echoes constitutes a linear distortion of the TV signal if one considers received base-band in relation to transmitted base-band. This notwithstanding the inherent non-linear character of modulation in general. It will be seen also that at RF level and during demodulation there is considerable interaction between the in-phase and quadrature channels in case of echoes but this does not affect the linearity of the process at base-band

level. The assumption here is that demodulation is done by synchronous detection [15][16][17]. In 2.2.6 we come back to this issue, showing heuristically that asynchronous (envelope) [18][19] and quasi synchronous [15] detection can lead to non-linear effects at base-band level.

We thus conclude that as far as echo cancellation is concerned we need not consider non-linear distortions. Of course these do indeed occur but they are a subject in its own right and will be considered beyond the scope of this thesis on the basis of the preceding paragraph.

From this we can conclude that echo cancellation in principle can be performed by linear, adaptive filtering and that suitable filtering will not eliminate echoes only, but also other linear distortions.

2.2.2 Echo Statistics

In practice it turns out that many echoes, even a whole continuum of echoes can occur which have widely different strength, sometimes of the same order of magnitude as the un-echoed signal. Our terminology in this context will be such that the strongest of the received signals will be referred to as the main signal regardless of the question whether it has been echoed or not. Echoes arriving prior to the main signal will be called pre-echoes, echoes arriving after the main signal will be called post echoes.

Statistical studies [20], test experiences [21] and private experiences of us show that in practice pre-echoes arrive up to 5 μ sec prior to the main signal and post-echoes arrive up to 45 μ sec after the main signal.

If the GCR has a duration of t_0 μsec the echoed GCR will thus have a duration of up to $t_0 + 50$ μsec of which the 50 μsec are unavoidable once one uses a GCR at all. Since a TV line has a duration of 64 μsec we thus see that the introduction of a GCR in the VBI roughly causes at least an echoed version of about one line duration. Now the allotment of line space to different services is done in the broadcast world on a per line basis and therefore we see that if the GCR is to be any longer than a few μsec we might as well give it a length of about one line duration. This will turn out to be sufficient in terms of the energy content required by signal to noise considerations.

2.2.3 Cross effects between GCR and the contents of neighboring lines

If we accept for the moment the idea that the GCR should be confined to one line in the VBI we are left with the simple situation that the GCR will be echoed onto the next line, while the information, if any, on the line preceding the GCR will be echoed onto the GCR. Since comparison in the receiver between received and stored GCR requires the received GCR to be uncompromised by information on neighboring lines one would expect that it becomes necessary to omit all information from these neighboring lines. We will show below (see 2.3.2) that this is not fully true: by properly cycling the sign of the transmitted GCR, proper choice of the lines to be used and proper averaging procedures these difficulties can be prevented to a considerable extent.

2.2.4 GCR and Channel Characterization

Obviously the comparison in the receiver of received and stored GCR must

serve in the first place to characterize the echoing channel. When discussing cancellation in detail in later chapters we will have to treat this issue in depth. A few things of primary importance can however be mentioned at the intuitive level here.

In the first place the channel must be properly characterized over its full bandwidth. For this two things are necessary: the GCR itself should have the full bandwidth of the base-band channel and the frequency characteristic of the channel should have sufficient magnitude and stability over the full frequency range. This is not always the case. Under certain circumstances a "notch" in the channel spectrum can exist [22][23] to such an extent that restoration would imply the introduction of noise if not instability. This can result in a situation where it is better to have no cancellation at all. Therefore the canceller has to be able to diagnose the situation in this respect and to shut itself down if the need arises (see 12.2).

In the second place the GCR itself should be as immune to noise as possible. Its amplitude, like the amplitude of all TV signals is subject to regulation. As we have said before, its duration is limited by availability of line space. So spectrum-wise as well as time-wise the available resources must be used as economically as possible. This points in the direction of a flat spectrum within the bandwidth and a constant amplitude within the duration of the GCR.

2.2.5 Sampling Requirements

As said before the use of adaptive filtering for echo cancellation implies, in view of today's technology, digital signal processing. Therefore a clocking system is needed to obtain for instance properly timed samples of the echoed GCR. We have already seen that the sampling frequency that we need is of the order of 10

MHz, corresponding to a sampling period of 100 n sec. Clearly the accuracy of the clock, and therefore of the Phase Locked Loop that controls it, must be a small fraction of the sampling period which brings us in the $< 10 n$ sec region. If the clock lacks accuracy the high frequency components of the sampled GCR tend to average out and consequently the echo cancellation process would boost these frequencies in the video signal even when there is no need for this at all from the point of view of echo cancellation.

2.2.6 A-synchronous and Quasi Synchronous Detection

As we stated before, linearity of the echo creation process at base-band level can be proven to exist if synchronous detection [16][17] is used at the demodulation stage. This is the main result of our considerations in chapter 4.

Here we would like to show in a heuristic way that a-synchronous (or amplitude) and quasi synchronous detection can introduce non-linearities.

In a-synchronous or envelope detection [17][19] we basically determine the amplitude of the received RF signal. Consider the relatively simple, and for TV not even realistic, case of Double Side-Band Amplitude Modulation. Here, before modulation the information carrying base-band signal $f(t)$ is multiplied with a modulation index m so that $mf(t) < 1$ at all times, after which 1 is added. This guarantees that the resulting base-band signal is positive. Then modulation takes place by multiplication with $\cos(\omega_0 t)$, where ω_0 is the RF carrier frequency. If there are no echoes we simply can determine the amplitude $1 + mf(t)$ of the RF signal in the receiver and derive $f(t)$ as described in chapter 4. Thus in this utterly simple situation the process from base-band to base-band is linear. This may

change when echoing occurs. A good example is the case, where $1 + mf(t)$ decreases from a high (positive) constant value V_1 to a small (positive) constant value V_2 at a certain point in time and where at the site of the receiver apart from the main signal an echo of strength $a < 1$ arrives with such a delay that (by coincidence) the phase of the RF carrier is reversed. Now for a time equal to the delay we will receive $(V_2 - aV_1)\cos(\omega_0 t)$ where $V_2 - aV_1$ can be negative, which we assume to be the case here. Before this period we would receive $(V_1 - aV_1)\cos(\omega_0 t)$ and after the period we would receive where $(V_2 - aV_2)\cos(\omega_0 t)$, where $V_1 - aV_1$ and $V_2 - aV_2$ are positive. Obviously these sign changes in the factor before $\cos(\omega_0 t)$ will not be detected by amplitude detection which would only yield the absolute value of that factor. Clearly this would constitute a grave nonlinearity, which obviously cannot be removed by linear processing.

In the case of quasi synchronous detection we have a very similar situation. It can be illustrated by using exactly the same example as in the above discussion of amplitude detection. In quasi synchronous detection [15][17] we take the incoming RF signal, amplify it very strongly, limit it in amplitude and multiply the incoming RF signal with this square wave version of the RF carrier as sensed at the receiver side. Thus in the period of time in which the factor before $\cos(\omega_0 t)$ is negative, we know that the carrier wave as perceived by the receiver has flipped sign. Therefore multiplication with $\cos(\omega_0 t)$ gives a negative result and so multiplication with the factor before the cosine (which is negative here) leads to a positive end result. In the period of time in which the factor before $\cos(\omega_0 t)$ is positive, we know that the carrier as perceived by the receiver has its original sign. Therefore multiplication with $\cos(\omega_0 t)$ gives a positive result and so multiplication with the factor before the cosine (which is positive here) leads to a positive end result. We therefore have essentially the same grave nonlinearity as in the previ-

ous case. The difference between the present case and the case of amplitude detection is that in the present case the output signal obviously is zero whenever the square wave that we used goes through zero.

These examples should have made clear that nonlinearities in DSB-AM already can, but do not necessarily occur unless we use synchronous detection. Since DSB-AM constitutes a part of VSB-AM the same holds for VSB-AM, which is commonly used in TV. Therefore if we want to stick to linear echo cancellation methods in the present work we will have to assume that the receiver uses synchronous detection.

2.3 Design

In what follows in this section we already mention those design principles that can be made clear even before we begin to discuss our actual design: we present some initial considerations for the GCR design, consider its proper cycling on a field to field basis and make a few remarks on the overall structure of a canceler.

In the following we confine ourselves to NTSC M standards and values unless stated otherwise.

2.3.1 Design Principles for the GCR

The details of the design of the family of GCRs that we proposed will be dis-

cussed in chapters 5 and 6. We would like to point out at this stage the most important ones in a qualitative way. In the first place there is a choice to be made on whether the GCR is to be a complex signal [24] or a real signal. This is not a trivial question since Vestigial Side-Band Modulation [11] involves Single Side-Band [25] components in the RF domain, so that we have to take into account what happens in both the in phase and quadrature channel. This might imply that the GCR has to have components in both channels i.e. that the GCR could be required to be complex. In fact complex GCRs have been proposed in the past [26].

As we can see from Chapters 4 and 5 this will turn out not to be the case, so that we can assume the GCR to be a real signal. This of course gives cost economy in the echo canceler at the receiver side.

We already stated that the GCR should cover the whole bandwidth of the channel and that it should be confined to one line of the VBI where it should have constant amplitude in the course of time in order to carry as much energy as possible. Furthermore it should be positioned in the middle of the dynamic range that is allowed to TV signals in order to ensure as much linearity as we can by escaping nonlinear effects that may come about in the transmission and receiving hardware. This makes it necessary to place the GCR on a pedestal that is situated midway between the black level at (-40 IRE) and the white level (100 IRE) and to give it an amplitude of 40 IRE. Here "IRE" means the universally accepted linear unit in terms of modulation depth to measure the relative strength of various parts of video signals in TV. The length of the GCR was chosen to be 35.5 μ sec and the pedestal itself occupies all the line space that is available after the 11.8 μ sec that has been reserved [3] for sync pulse and color burst. In the USA line 19 of the VBI has to be used for placing the GCR.

2.3.2 Cycling Considerations

In order to reduce the effects of spilling information from one line to another by echoing (see 2.2.3) we have to use a proper cycling scheme. Such a cycling scheme is different for different TV systems. In NTSC [3] it is based on the fact that NTSC has an interlaced [3][27] scanning structure, where each picture (frame) of 525 lines is scanned by first scanning the odd numbered lines (comprising one "field") and then scanning the even numbered lines (comprising the other "field"). The color burst changes sign every field [3][28] and the sync pulse does not change sign. Furthermore the content of the adjacent lines 18 and 20 needs to be considered. Line 18 contains information [29][30][31] that varies from field to field, line 20 contains stationary information except for its later part [29][30] where the occurrence of an echoed part of the GCR is not very likely.

The following way [31] of cycling the sign of the GCR at the transmitter during eight consecutive fields, while leaving the pedestal constant, was used:

Field number:	1	2	3	4	5	6	7	8
GCR sign.	+	-	+	-	-	+	-	+

The following table shows, that if we apply the same sign changes to the GCR and its neighboring lines in the receiver we cancel constant information on all three lines, such as the pedestal and syncs, as well as the color bursts.

Field number	1	2	3	4	5	6	7	8
--------------	---	---	---	---	---	---	---	---

GCR as sent	+	-	+	-	-	+	-	+	
GCR as received	+	+	+	+	+	+	+	+	sum not zero
Constants sent	+	+	+	+	+	+	+	+	
Constants received	+	-	+	-	-	+	-	+	sum zero
Burst as sent	+	-	+	-	+	-	+	-	
Burst as received	+	+	+	+	-	-	-	-	sum zero

On top of this, randomly varying information on lines 18 and 20 tends to be averaged out. This averaging can be improved (at the cost of slower cancellation) by extending the cycle to a multiple of 8 fields.

2.3.3 Hardware and Software in the Echo Canceled

As to hardware and software in the echo canceler, what is in principle needed is means to properly sample lines 18, 19 and 20 of the VBI with sufficient temporal accuracy to guarantee proper superposition of the results of 8 consecutive fields of the GCR cycle, analog to digital conversion of the results as well as of the incoming video signal, a system to characterize the channel using a comparison of received and stored GCR, a microprocessor to do the comparison and to calculate from it the proper programming of the adaptive linear filter system that will cancel the echoes from the TV signal, the filter system itself and digital to analog conversion of the processed video signal for processing in the conventional part of

the receiver. All of this is summarized in Figure 2.3.3.F1. We comment below on those items mentioned above where conventional means were improved on in our system. These are channel characterization, the filter system and the programming of a FIR filter for cancellation of pre-echoes.

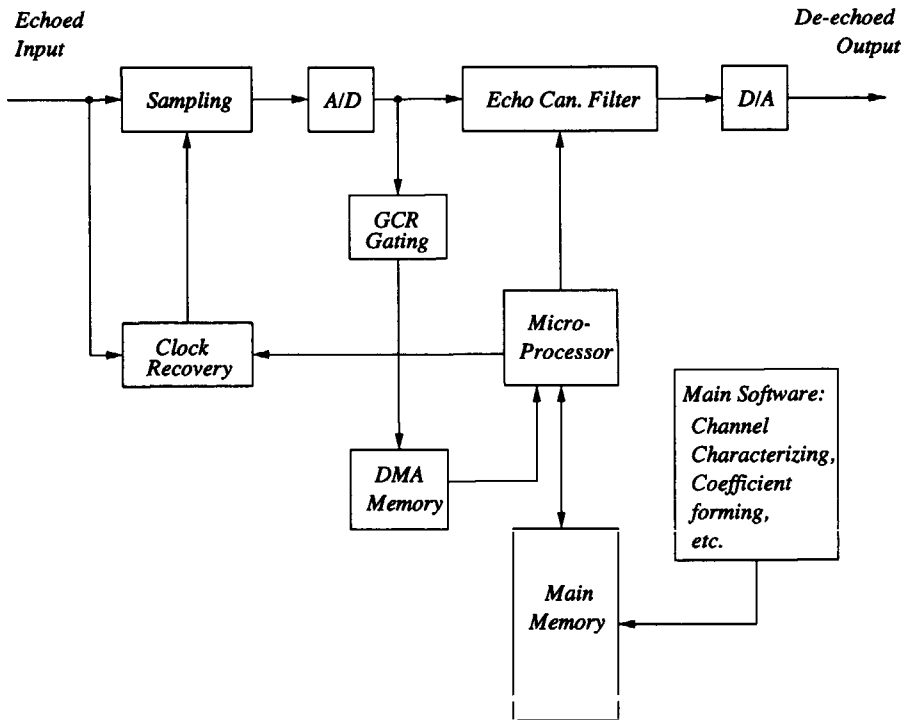


Figure 2.3.3.F1

A Simple Block Diagram of the Echo Canceling System

2.3.3.1 Channel Characterization

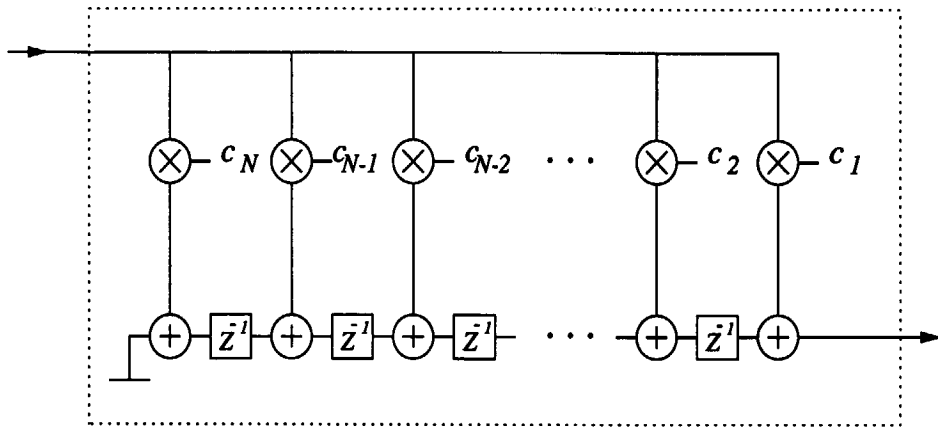
We show in chapter 5 that if the GCR is a real signal with a flat spectral density over the whole band, the correlation [32] of the received GCR and the stored GCR is identical with the impulse response function of the channel. Furthermore we indeed succeeded in the design of our family of GCRs to have such a flat spectral distribution (see Figure 6.2.F3). We thus were able to use this correlation method and did so in most of our work. It was one of the tasks, as the reader might expect, done by the microcomputer mentioned above.

2.3.3.2 The Filter System

The filter system required the implementation of a custom Integrated Circuit for reasons that we will now describe.

Two types of digital linear filters are commonly used, known as Finite Impulse Response (FIR) [33] and Infinite Impulse Response (IIR) [33] filters respectively.

A FIR filter, see Fig 2.3.3.2F1, consists of an input line from which the signal is led through a series of parallel (programmable or fixed) multipliers to a series of adders, connected by delay elements with equal delays. One end of this series is connected to ground, the other is the output.

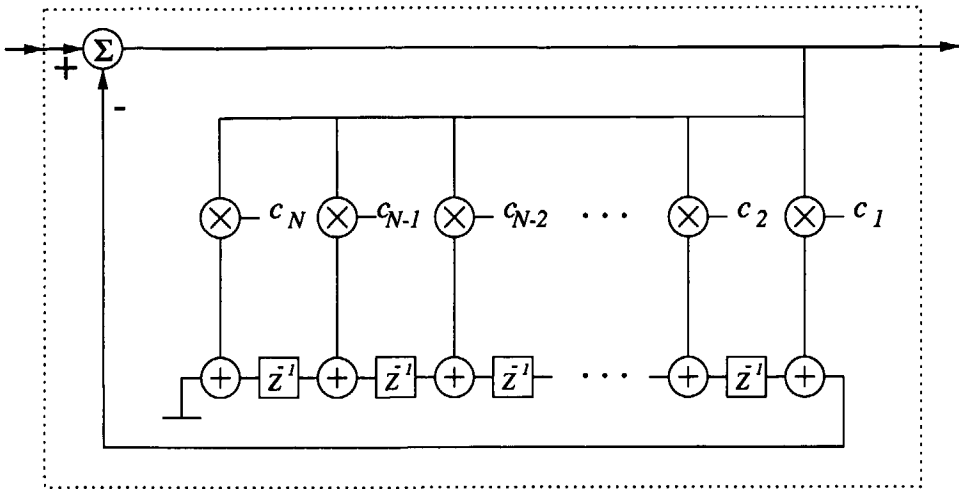


FIR

Figure 2.3.3.2.F1
A Typical Structure of a FIR Filter

An IIR filter consists of a recursive structure as shown in Fig 2.3.3.2.F2; it consists of an adder at the input that adds a FIR filtered version of the output with reversed sign to the input, the sum itself being the output. It is easy to see [33] that if the FIR filter has a frequency response $H(\omega)$ the total IIR filter has a frequency response

$$H_{IIR}(\omega) = \frac{1}{[1 + H(\omega)]} \quad (1.3.3.2.E1)$$



IIR

Figure 2.3.3.2.F2
A Typical Structure of a IIR Filter

Clearly, because of the direct connection of the input of an IIR filter with its output, it creates unprocessed output as soon there is input, and therefore it can never be used to remove pre-echoes. On the other hand we see from 1.3.3.2.E1 that, apart from stability issues, it can be used to neutralize the effects of a filter with frequency response $1 + H(\omega)$, which makes it very attractive for echo cancellation purposes.

We conclude [34] that pre-echoes have to be removed by a FIR filter for which the proper programming method still has to be discussed, and that possibly the post echoes can be canceled by an IIR that neutralizes the combined effects of the channel and the FIR filter that removes the pre-echoes. This would require a

concatenation of channel model and pre-echo FIR for the construction of its feedback loop.

There is one more detail which is relevant enough to be mentioned here. This is the fact that a FIR filter when used to cancel pre-echoes tends to do this at the cost of creating a pre-echo of reduced strength at a greater (negative) delay than the original ones.

To understand this consider a delta pulse with a pre-echo at time $-t_0$ of strength a . The way it is to be canceled by the FIR filter is to give the first "tap" a weight $-a$ and to give the tap at delay $+t_0$ a weight 1, all other taps having weight 0. The action of the filter now can easily be visualized pictorially by convolving the signal as depicted in Fig. 2.3.3.2.F3.a with the filter as depicted in Fig. 2.3.3.2.F3.b, leading to an output as depicted in Fig. 2.3.3.2.F3.c.

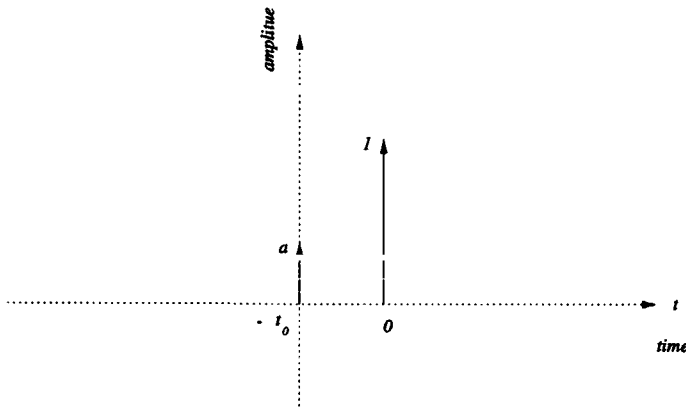


Figure 2.3.3.2.F3.a

A Symbolic Example of an Input Signal
Having an Pre-echo with Amplitude "a"

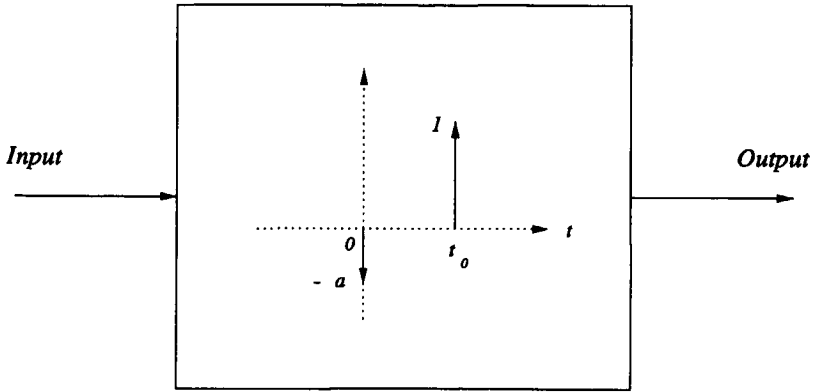


Figure 2.3.3.2.F3.b

A Symbolic Example of a FIR Echo-canceling
Filter with its Proper Coefficients

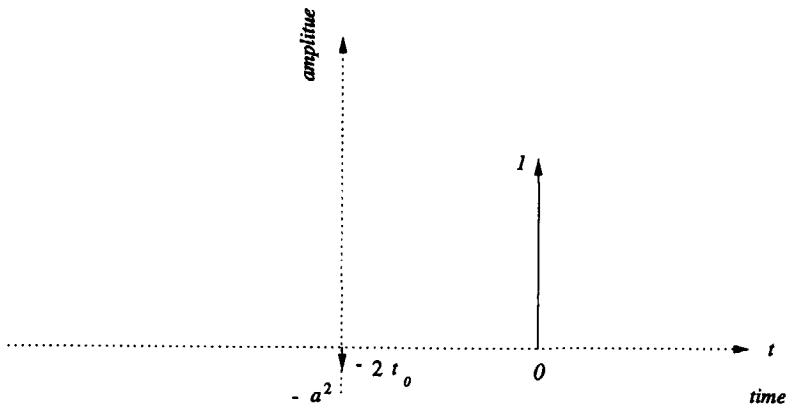


Figure 2.3.3.2.F3.c

The Output Signal of the Symbolic FIR Filter
with the First Order Pre-echo Canceled

The equivalent story in words is that when the pre-echo arrives at the input it is multiplied by tap 1 with $-a$ (and therefore reduced by a factor $-a$) and immediately delivered to the output. After a time t_0 the delta pulse itself arrives at the input while the pre-echo itself arrives at the output from the tap at delay t_0 . This causes an output of $-a$ times the delta pulse plus the pre-echo which is $+a$ times the weight 1. This output of course equals zero. Again after a second period t_0 the main delta pulse arrives at the output from the tap at t_0 , multiplied by 1. After this nothing is output any more. Thus the total output consists of the main pulse and a "pre-echo" at $-2 t_0$ of reduced strength. This new defect can be removed in the same way, producing a still weaker offspring at $-4 t_0$, etc. So we conclude that FIR filters that cancel pre-echoes perfectly have to be infinitely long. Obviously a trade-off has to be made. In any case it is a good thing that pre-echoes happen only at short times before the main signal arrives. In practice we use the FIR filter to remove the pre-echoes and only the very near-by post-echoes. All the other echoes have to be removed by the IIR filter. In its feedback loop we need a concatenation of the channel response and the pre-echo FIR filter as we saw earlier. This requires a convolution of the channel impulse response function and the pre-echo filter, as shown in Fig. 2.3.3.2.F4. Thus the FIR filter in the IIR loop gets shorter if the pre-echo filter is made shorter.

From the above paragraph we see that it may be necessary to switch filter resources from the IIR to the FIR filter or vice versa depending on the echoing situation. This is one of the special features [35] of the custom IC mentioned in the beginning of this section.

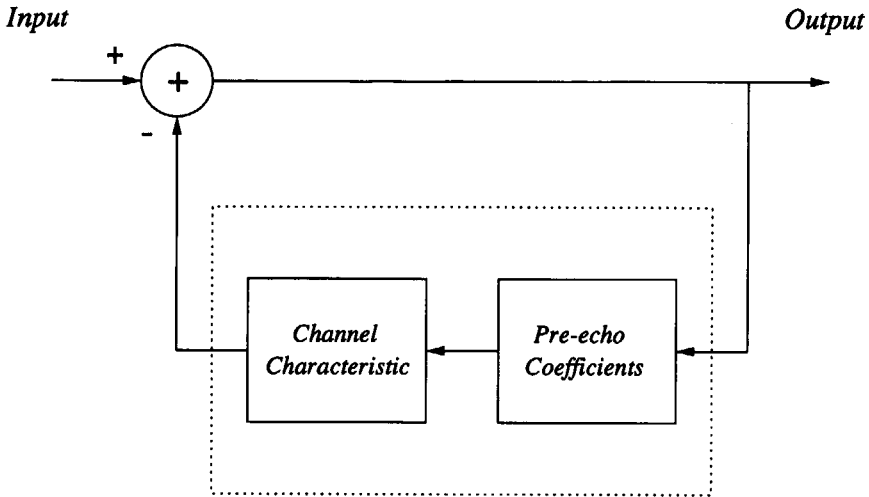


Figure 2.3.3.2.F4

A Symbolic Equivalency of the IIR Coefficients in an Echo Canceler

2.3.3.3 Programming the FIR Filter for Pre-echoes

The programming of the FIR filter for pre-echo removal is not trivial. We have to take recourse to adaptive methods [36] in order to create successively more accurate sets of tap values, which requires extensive processing based on correlation of the channel impulse response function with the error in the signal appearing at the output of the FIR filter. The details of this process [36] are given in Chapter 9. Obviously all processing has to be done by the micro processor.

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CHAPTER 3

PRELIMINARY INTRODUCTION TO ECHOES

This chapter is meant to be an introduction to echoes that may appear during the transmission of waves in a linear medium in general and in a TV broadcast in particular. They are always caused by some inhomogeneity in the medium that either provides an additional transmission path or disrupts the continuity of an existing transmission path.

The chapter merely is meant to place the next detailed chapter on echo creation in real TV channels in context and to be an introduction to it.

3.1 Echoes in Natural Channels (Linear, no Dispersion, Infinite Bandwidth)

In the absence of echoes, in a channel without dispersion and with infinite bandwidth, the response at the receiving end to a delta pulse at the transmitter is a delta function again: the impulse response function $h(t)$ in this case is a $\delta(t)$ pulse. Therefore the spectral transmission characteristic $H(\omega)$ is the Fourier transform of a $\delta(t)$ pulse i.e. a constant as a function of frequency.

In the case of an actual TV broadcast the assumption of infinite bandwidth is not valid since we use for separate transmissions separate channels with a bandwidth that is in practice about 6 MHz. However inside such a channel the dispersion can safely be assumed to be zero.

In order to understand qualitatively (for the time being) what happens in that case we proceed with considering first a natural channel, defined as a linear channel without dispersion and of infinite bandwidth and with discrete echoes only. Discrete echoes are defined here as echoes that each evoke an additional $\delta(t)$ function response that has a mostly positive (but sometimes negative) delay as compared with the non-echoed response. Such echoes are produced by reflections against sharp discontinuities such as buildings. Obviously the impulse response function in this case is a series of $\delta(t)$ functions at different delay times Δt_i and at different amplitudes a_i :

$$h(t) = \sum_{i=-M}^{N-M-1} a_i \delta(t - \Delta t_i) \quad (3.1.E1)$$

where the term with index 0 represents the direct signal: $a_0 = 1$

Since reflection is a passive process, energy is never increased by it and therefore all a_i have absolute values that are smaller than or equal to 1 for $i \neq 0$.

There are three remarks here that already at this early stage of the treatment deserve special emphasis:

1. Even in the presence of a single echo equation 3.1.E1 has two terms and for that reason the Fourier transform is not "flat" i.e. is not a constant as a function of frequency:

$$h(t) = a_0\delta(t) + a_1\delta(t - \Delta t_1) \quad (3.1.E2)$$

$$H(\omega) = a_0 + a_1 e^{-j\omega\Delta t_1} \quad (3.1.E3)$$

If the absolute value of a_1 equals a_0 the channel acts like comb filter and we have zero transmission at certain frequencies. This is a primitive example of the "notches" [1][2] in a real channel that have cost us some grief in the actual design of an echo canceller as described in chapter 12.

2. The case of diffuse reflections, for example in the case of mountainous landscapes, can be treated by simply changing the sum in 3.1.E1 into an integral, thus increasing the number of echoes to infinity.
3. TV transmission is done as broadcast, i.e. by transmission through free space, and also through cable. In the latter case reflection against buildings and mountains does of course not occur, but instead there are reflections because of impedance mismatches [3][4] between the beginning and any particular endpoint or branching in the cable system which cause reflections as well. Since in cables we have considerable attenuation, only the echoes with smaller delay will influence the received sig-

nal in an observable way. Here of course it is assumed that at the cable head-end [5] no echoes are present. If they are they should be canceled right there.

3.2 Echo Modeling for a Natural Channel

For the "natural" channel that we considered in the previous section, linear with no dispersion, only discrete reflections and infinite bandwidth, we will now briefly indicate one way in which the channel can be characterized. This makes sense because all results to be obtained later with various types of modulation and subsequent transmission through band-limited channels can be translated to the elementary case treated here. Perhaps superfluously, we note that echo cancellation obviously presupposes that the channel from the transmitter to the receiver in question has been properly characterized.

If the input to the channel is $x(t)$ with Fourier transform $X(\omega)$ and the output is $y(t)$ with Fourier transform $Y(\omega)$ then:

$$y(t) = x(t) * h(t) \quad (3.2.E1)$$

and

$$Y(\omega) = X(\omega) H(\omega) \quad (3.2.E2)$$

Therefore if both input and output are known the channel characteristic can in principle be obtained. Evidently the output is known from the receiving process. The input as a whole is not known to the receiver, but in the case of TV we can standardize a periodically recurring part of the input, to be sent during the vertical blanking interval and store that part permanently in a memory in the receiver, thus enabling it to compare input and output and to construct $h(t)$ or $H(\omega)$ using 3.2.E1 or 3.2.E2. The standard signal, commonly called Ghost Cancellation Reference Signal (GCR), has to be carefully designed in order to meet the numerous requirements that it has to satisfy. In later chapters these requirements, the subsequent design process and its final result will be treated at length.

3.3 Echoes in a Real Channel as Considered from Base Band to Base Band, Taking Modulation into Account

As we said before the base-band signals of separate TV stations are modulated into separate channels of limited bandwidth in practice. This means that these base-band signals are shifted [6][7] in frequency into the RF region in a way the details of which depend on the modulation method. This is vestigial side-band modulation [6][7] in regular TV, which as we will see is quite complicated from the echo point of view. In fact the echoes are introduced at the RF level so that an echo delay of Δt_i results in a phase difference involving the product of Δt_i and the RF frequency rather than the base-band frequency. Furthermore the vestigial side-band modulation process causes part of the signal to be Hilbert transformed [8][9] as we will see later. One important characteristic of this phenomenon is that the impulse response of a Hilbert transformed $\delta(t)$ function is not a $\delta(t)$ function

but rather a lopsided version of it. In sum after demodulation at the receiver we have in the real world totally different echo patterns from different channels because they all have a different RF frequency and we no longer necessarily have just a series of δ functions in response to a $\delta(t)$ function, even if the echoes are discrete.

In spite of all these complications it remains true that the whole process from base-band at the sending end, through modulation, echoing and demodulation to base-band at the receiver end remains linear if we use synchronous detection [10][11] in the demodulation process and consequently a comparison of base-band output with base-band input gives us a "channel" characteristic $h(t)$ or $H(\omega)$ with the same meaning as in the case of the natural channel and therefore echo cancellation can be performed in the same way on the basis of this characteristic. The only difference is that the impulse response function of this "base-band equivalent" channel cannot be naively interpreted anymore by the human observer in terms of a series of (band-limited) $\delta(t)$ functions.

Needless to say that the quantitative mathematics of all this will be presented in considerable detail later in this text. The foregoing was only meant as a description in advance of things to come in order to facilitate further reading.

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CHAPTER 4

ECHOES IN TELEVISION CHANNELS

In this chapter we will describe in a rigorous way the process through which echoes in television are created. As stated already in the previous chapter this happens at RF level. For this reason everything depends on the modulation and demodulation technique that is employed. In practice this is Vestigial Side-Band AM in terrestrial and cable television [1][2], which contains Double Side-Band [1][2] and Single Side-Band AM [1][2] as elements.

We therefore treat all three modulation methods in the order of increasing complexity in order to show clearly where all the effects come from. One of the main results will be that at baseband level the relation between input and output of the total chain is linear, although the effects of the various processes at RF level are clearly visible in the equations.

In later chapters we will use this result in connection with channel characterization and echo cancellation.

4.1 Echoes in Double-side-band Amplitude Modulation (DSB AM) and Demodulation

4.1.1 DSB AM Modulation

DSB AM modulation [1] is the modulation scheme that can most easily be understood among all AM schemes. Furthermore the actual modulation scheme in terrestrial television, i.e. vestigial-side-band modulation (VSB AM) has DSB AM as an essential element, together with single-sideband-modulation (SSB AM). For these reasons we will start our analyses with DSB AM.

Amplitude Modulation [1] of course historically means the modulation of the amplitude of a carrier wave. Because of this it is not surprising that the demodulation method that was historically used to recover the modulating signal was the monitoring of the amplitude of the carrier wave, a process that became known as amplitude detection [3]. In connection with this it was considered useful at the time to introduce a so-called modulation index [1], to be defined below. In our notation we will retain this index although in the context of the more modern synchronous detection [3] this is not necessary. In fact there are still TV sets on the market in which amplitude detection is used. As will become clear in the course of this book, Ghost Cancellation can only be adequately performed by using synchronous detection.

Let us assume that at the modulator we have a carrier wave with an amplitude A and a frequency ω_0 which we write as $A \cos(\omega_0 t)$. We modulate an information carrying signal $f(t)$ on it by multiplying $A \cos(\omega_0 t)$ with $1 + mf(t)$, where the inclusion of the term 1 indicates that we are treating the case of an un-suppressed carrier [1], and where m is the modulation index mentioned above.

Here the normalization convention [1] is that $-1 < f(t) < 1$ and $0 < m < 1$. Consequently $1 + mf(t)$ cannot be negative and therefore the carrier wave is never multiplied with a negative number which permits us to monitor the amplitude only (instead of amplitude and phase together) for detection without ambiguity. We will refer to $1 + mf(t)$ as $g(t)$ sometimes and to the total signal $A[1 + mf(t)] \cos(\omega_0 t)$ as $s_d(t)$, where the subscript d refers to DSB:

$$s_d(t) = A[1 + mf(t)]\cos(\omega_0 t) \quad (4.1.1.E1)$$

$$= g(t)A\cos(\omega_0 t) .$$

If the Fourier transforms of $s_d(t)$ and $f(t)$ are $S_d(\omega)$ and $F(\omega)$ respectively and if the Fourier transform of $g(t)$ is $G(\omega)$, we have in the frequency domain, i.e. as the Fourier transform of 4.1.1.E1:

$$S_d(\omega) = A\pi \delta(\omega + \omega_0) + A\pi \delta(\omega - \omega_0) + A\frac{1}{2} mF(\omega + \omega_0) + A\frac{1}{2} mF(\omega - \omega_0)$$

$$= A\frac{1}{2} G(\omega + \omega_0) + A\frac{1}{2} G(\omega - \omega_0) \quad (4.1.1.E2).$$

$S_d(\omega)$ is shown in comparison with $F(\omega)$ in Fig 4.1.1.F1 and Fig 4.1.1.F2:

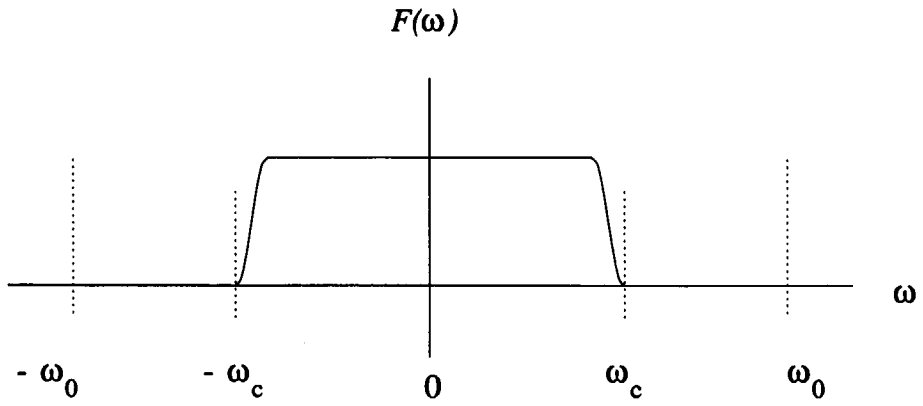


Figure 4.1.1.F1
The spectrum of the Base-band Signal $f(t)$

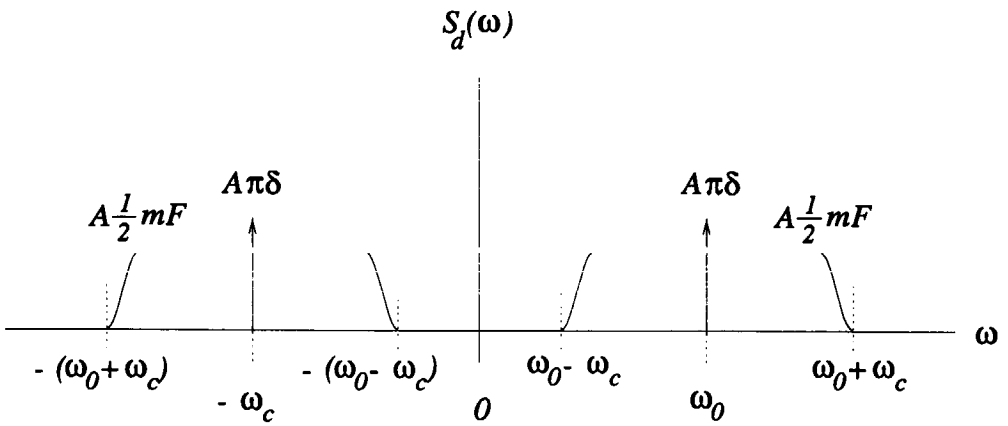


Figure 4.1.1.F2
The Spectrum of a DSB-AM Modulated Signal

We see there the well-known result [1][2] that the spectrum of a DSB AM modulated signal $f(t)$ with unsuppressed carrier consists of two copies of the original spectrum of $f(t)$ translated by ω_0 and $-\omega_0$ respectively plus two delta functions at ω_0 and $-\omega_0$. The latter of course represent the unsuppressed carrier. If the maximum frequency in $F(\omega)$ ω_c exceeds ω_0 the two spectra will overlap. In that case they will cause aliasing and cannot be separated by filtering and as we shall see in the following section synchronous detection will fail.

4.1.2 DSB AM Demodulation

We have already mentioned in 4.1.1 that amplitude detection can be used for double-side-band demodulation. We also mentioned that this method will turn out to be unsuitable in the case of ghost cancellation. Instead we have to use synchronous detection [3]. In its most elementary form, which however is applicable to the unsuppressed carrier case, synchronous detection presupposes that at the receiver side the carrier wave has been recovered with accurate frequency and phase. Let it suffice to say here that this is technically possible.

It thus is possible to multiply the received signal $AB[1 + mf(t)]\cos(\omega_0 t)$ with $2C\cos(\omega_0 t)$ again, which results in $D[1 + mf(t)]2\cos(\omega_0 t)\cos(\omega_0 t)$, where A, B and C have been condensed in one constant D, which is unknown due to uncertainty in B due to attenuation and amplification in the signal path.

We can now write

$$r_d(t) = D [1 + mf(t)] 2\cos(\omega_0 t) \cos(\omega_c t) \quad (4.1.2.E1)$$

$$= D [1 + mf(t)] [1 + \cos(2\omega_0 t)]$$

$$= Dg(t) [1 + \cos(2\omega_0 t)]$$

$$= Dg(t) \left[1 + \frac{1}{2} e^{j2\omega_0 t} + \frac{1}{2} e^{-j2\omega_0 t} \right].$$

The three terms in this equation have spectra that extend from $-\omega_c$ to $+\omega_c$, $2\omega_0 - \omega_c$ to $2\omega_0 + \omega_c$ and $-2\omega_0 - \omega_c$ to $-2\omega_0 + \omega_c$. Obviously they can be separated only if $\omega_0 > \omega_c$.

Low pass filtering then provides us with the demodulated signal $\tilde{r}_d(t)$ (the ~ here associated with the function stands for the low-pass filtered version of that function):

$$\tilde{r}_d(t) = D [1 + mf(t)] \quad (4.1.2.E2)$$

$$= Dg(t).$$

This is still not the information carrying signal $f(t)$. If $f(t)$ would have no dc component, as would be the case in audio, the average of $\tilde{r}_d(t)$ over time would be D . Division of $\tilde{r}_d(t)$ by D would yield $1 + mf(t)$ and m would be determined

by the standards that transmitter have to adhere to. So $f(t)$ would be known. In the case of video there is a dc component, representing the average brightness of the screen. So now $\tilde{r}_d(t)$ contains two contributions to its dc component; 1 and m times the average of $f(t)$, and at the receiver there is no way to distinguish between them using the equations derived above. The proper strength of the time varying part of $f(t)$ cannot even be determined since it occurs in $\tilde{r}_d(t)$ in a triple product $Dmf(t)$ of which only m is known.

The way out as used in the industry [4] is to use the convention in most systems and a related one in the others that during the blanking intervals $f(t)$ is sent as black: depending on the television system the minimum or the maximum video voltage apart from the line or field sync signal. Since m is also fixed by the standards D can then be determined from $\tilde{r}_d(t)$ and thus $f(t)$ also. That would complete the detection in this case.

4.1.3 Echoes in the DSB AM Modulated signal

Now assume that a directly transmitted signal as well as a reflected signal (echo) arrive at the receiver. In order to consider the impact of this we have to realize that the actual echo is created while the signal travels at RF: it has not yet been demodulated yet in the receiver. The RF echo can be described by an attenuated and delayed version of $g(t) = [1 + mf(t)]\cos(\omega_0 t)$ where the delay has to be expressed in $f(t)$ as well as in $\cos(\omega_0 t)$. Therefore the echo is

$$e_d(t) = as_d(t-\Delta t) \tag{4.1.3.E1}$$

$$= ag(t-\Delta t)\cos[\omega_0(t-\Delta t)]$$

$$= a \cos(\omega_0\Delta t) g(t-\Delta t)\cos(\omega_0 t) + a \sin(\omega_0\Delta t) g(t-\Delta t)\sin(\omega_0 t) .$$

If there are more echoes we have to sum them. The total received signal $\hat{s}(t)$ will contain the directly transmitted signal and all the echoes:

$$\hat{s}_d(t) = s_d(t) + e_d(t) \tag{4.1.3.E2}$$

$$= g(t)\cos(\omega_0 t)$$

$$+ \sum_{i=0}^{n-1} a_i \cos(\omega_0\Delta t_i) g(t-\Delta t_i)\cos(\omega_0 t)$$

$$+ \sum_{i=0}^{n-1} a_i \sin(\omega_0\Delta t_i) g(t-\Delta t_i)\sin(\omega_0 t).$$

The first sum in 4.1.3.E2 is called the "in phase" part of the echoes in the RF domain since it contains $\cos(\omega_0 t)$. The second sum is called the "quadrature" part of the echoes because it contains $\sin(\omega_0 t)$.

4.1.4 Echoes in DSB AM Demodulation

As noted above we have in 4.1.3.E2 terms with $\cos(\omega_0 t)$ as well as a term

with $\sin(\omega_0 t)$. This is easily understandable because the reflection causes a delay and thus a phase shift at the RF level. As a consequence there is no special reason to use just $\cos(\omega_0 t)$ for demodulation: it would seem that we can or even have to multiply instead with a phase shifted version: $\cos(\omega_0 t + \phi)$. However the proper phase shift would depend on the strength a_i and on the delay Δt_i of the echoes, which before correction for the echo are unknown. Furthermore at this stage of the reasoning it is not clear whether a "proper" phase shift can be defined at all since from 4.1.3.E2 it can be seen that the RF phase shift of the received signal $\hat{s}_d(t)$ is not even constant but time dependent. Therefore we can only investigate what happens if we do the demodulation by multiplication by $2\cos(\omega_0 t + \phi)$ where the phase angle ϕ is arbitrary. Executing this program will involve complex equations as we shall see shortly, but fortunately most of the terms have double frequency and therefore vanish in the end by low pass filtering. We write and modify the expressions for the demodulated but not yet low pass filtered signal as follows:

$$r_d(t) = \hat{s}_d(t)2\cos(\omega_0 t + \phi) \quad (4.1.4.E3)$$

$$= \hat{s}_d(t)2\cos(\omega_0 t)\cos\phi - \hat{s}_d(t)2\sin(\omega_0 t)\sin\phi$$

$$= g(t)\cos(\omega_0 t)2\cos(\omega_0 t)\cos\phi$$

$$+ \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) g(t - \Delta t_i) \cos(\omega_0 t) 2\cos(\omega_0 t) \cos\phi$$

$$+ \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) g(t - \Delta t_i) \sin(\omega_0 t) 2\cos(\omega_0 t) \cos\phi$$

$$- g(t) \cos(\omega_0 t) 2 \sin(\omega_0 t) \sin \phi$$

$$- \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) g(t - \Delta t_i) \cos(\omega_0 t) 2 \sin(\omega_0 t) \sin \phi$$

$$- \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) g(t - \Delta t_i) \sin(\omega_0 t) 2 \sin(\omega_0 t) \sin \phi$$

$$= \cos \phi g(t) + \cos \phi g(t) \cos(2\omega_0 t)$$

$$+ \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) \cos \phi g(t - \Delta t_i)$$

$$+ \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) \cos \phi g(t - \Delta t_i) \cos(2\omega_0 t)$$

$$+ \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) \cos \phi g(t - \Delta t_i) \sin(2\omega_0 t)$$

$$- \sin \phi g(t) \sin(2\omega_0 t)$$

$$- \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) \sin \phi g(t - \Delta t_i) \sin(2\omega_0 t)$$

$$- \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) \sin \phi g(t - \Delta t_i)$$

$$+ \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) \sin \phi g(t - \Delta t_i) \cos(2\omega_0 t) .$$

Low pass filtering in order to remove the double frequency components delivers a relatively simple result $\tilde{r}_d(t)$, as promised in the above:

$$\tilde{r}_d(t) = \cos \phi g(t) \tag{4.1.4.E4}$$

$$+ \sum_{i=0}^{n-1} a_i \cos(\omega_0 \Delta t_i) \cos \phi g(t - \Delta t_i)$$

$$- \sum_{i=0}^{n-1} a_i \sin(\omega_0 \Delta t_i) \sin \phi g(t - \Delta t_i) .$$

This equation shows that the direct signal acquired a multiplicative factor $\cos \phi$ indicating that we will not even detect a direct signal if $\cos \phi = 0$, i.e. if ϕ is an odd positive or negative multiple of $\frac{\pi}{2}$. As will be discussed in 13.6 it is essential for proper echo cancellation that the main signal as demodulated in the receiver is as strong as possible compared to the echoes. For that reason we have to choose $\cos \phi$ as close to 1 as possible. This can be done [3] in most TV systems by having a phase locked loop in the receiver that locks onto the RF carrier wave only in those (short) periods in time in which the sync tips of the TV signal are received, the reason for this is that these sync tips have a video voltage that is maximal so that at the tips the video signal is relatively unsusceptible to echoes of other parts of the signal.

We note that 4.1.4.E4 contains as its third term a quadrature component which will turn out to be of high importance in the later parts of this chapter. Since this component is proportional to $\sin\phi$ it vanishes if $\cos\phi = 1$. So for DSB AM we can state that for $\cos\phi = 1$ no traces of events in the quadrature channel show up in the end result, even when echoes occur. We will see later that with SSB AM this is no longer the case: SSB AM produces a quadrature component in the modulation stage already, which in case of echoes gets shifted back into the in-phase channel.

The main conclusion from all this is that for DSB-AM the base-band response is linear in $g(t)$.

4.2 Echoes in Single-side-band Amplitude Modulation (SSB AM) and Demodulation

4.2.1 SSB AM Modulation

We will discuss SSB AM [1][2] on the basis of the spectrum that ensues after DSB AM modulation has occurred. As we saw in fig 4.1.1.F2 in that case we end up with a spectrum consisting of two $\delta(\omega)$ functions at frequencies $+\omega_0$ and $-\omega_0$, around which there is a copy of the spectrum of the original modulating signal $g(t)$, multiplied with the modulation index m . The delta functions represent the unsuppressed carrier wave. We see that the bandwidth that the DSB AM signal occupies is twice the bandwidth of the original signal. SSB AM attempts to remove this inefficiency by removing the lower sidebands of the two components of the DSB AM signal. Here "lower" means lower as to the absolute value of the

frequency.

This in general produces a complication for the dc part of the modulating signal, which we can overlook in the case of television, since television is actually sent in vestigial sideband modulation (see 4.3) where this difficulty does not occur. Also VSB AM has an unsuppressed carrier in its DSB part so that we need not account for it in the SSB part either.

It therefore suffices for the present treatment of SSB AM to assume a suppressed carrier which permits us to drop the modulation index m from the equations, and to assume the absence of dc.

The formal basis then for the discussion is a version of 4.1.1.E2 from which the $\delta(\omega)$ functions and m have been removed so that $g(t) \equiv f(t)$ and $G(\omega) \equiv F(\omega)$, representing the spectrum after DSB AM with suppressed carrier † :

$$S_d(\omega) = \frac{1}{2}G(\omega - \omega_0) + \frac{1}{2}G(\omega + \omega_0) \quad (4.2.1.E1).$$

We split this expression [5] into four parts representing respectively:

The lower sideband for positive frequencies.

† In the following discussion, the amplitude constants A , B , C and D from the previous sections become not important and therefore will not be introduced into the following equations for the sake of simplicity, unless specified otherwise.

The upper sideband for positive frequencies.

The lower sideband for negative frequencies.

The upper sideband for negative frequencies.

This yields:

$$S_d(\omega) = \frac{1}{2}G(\omega - \omega_0)_{0 < \omega < \omega_0} + \frac{1}{2}G(\omega - \omega_0)_{\omega > \omega_0} \quad (4.2.1.E2)$$
$$+ \frac{1}{2}G(\omega + \omega_0)_{0 > \omega > -\omega_0} + \frac{1}{2}G(\omega + \omega_0)_{\omega < -\omega_0}.$$

We remove from this the lower side bands by adding a version where the lower sidebands have been given a minus sign:

$$T(\omega) = -\frac{1}{2}G(\omega - \omega_0)_{0 < \omega < \omega_0} + \frac{1}{2}G(\omega - \omega_0)_{\omega > \omega_0} \quad (4.2.1.E3)$$
$$- \frac{1}{2}G(\omega + \omega_0)_{0 > \omega > -\omega_0} + \frac{1}{2}G(\omega + \omega_0)_{\omega < -\omega_0}.$$

A pictorial representation of this procedure is given in Fig.4.2.1.F1.

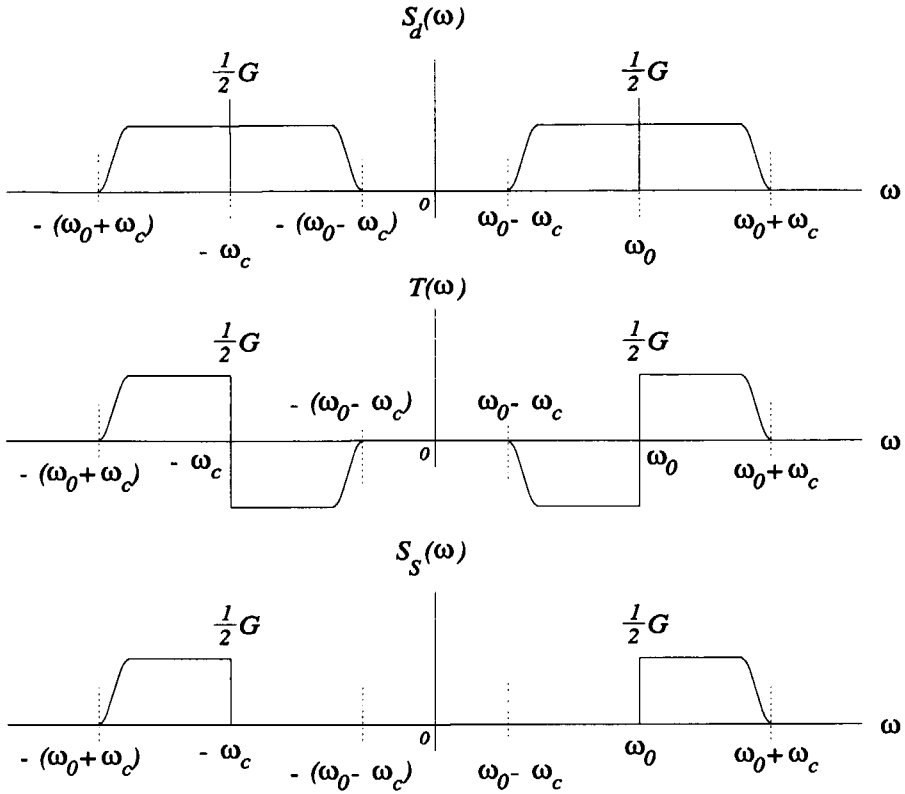


Figure 4.2.1.F1

The Relationship between $S_d(\omega)$, $T(\omega)$ and $S_s(\omega)$

To maintain the spectrum amplitude of the SSB AM signal in alignment with that of the DSB AM one, the SSB AM signal $S_s(\omega)$ can be derived from the sum of $\frac{1}{2} S_d(\omega)$ and $\frac{1}{2} T(\omega)$, which can be written as:

$$S_s(\omega) = \frac{1}{2} S_d(\omega) + \frac{1}{2} T(\omega) \quad (4.2.1.E4)$$

$$\begin{aligned}
&= \frac{1}{4}G(\omega - \omega_0) + \frac{1}{4}G(\omega - \omega_0) \operatorname{sign}(\omega - \omega_0) \\
&\quad + \frac{1}{4}G(\omega + \omega_0) - \frac{1}{4}G(\omega + \omega_0) \operatorname{sign}(\omega + \omega_0) \\
&= \frac{1}{2}G(\omega - \omega_0)_{\omega > \omega_0} + \frac{1}{2}G(\omega + \omega_0)_{\omega < -\omega_0} .
\end{aligned}$$

Fourier transform of this leads to the proper time domain representation of the suppressed carrier SSB AM signal:

$$\begin{aligned}
s_s(t) &= \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{4} G(\omega - \omega_0) e^{j\omega t} d\omega + \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{4} G(\omega + \omega_0) e^{j\omega t} d\omega \quad (4.2.1.E5) \\
&\quad + \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{4} G(\omega - \omega_0) \operatorname{sign}(\omega - \omega_0) e^{j\omega t} d\omega \\
&\quad - \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{4} G(\omega + \omega_0) \operatorname{sign}(\omega + \omega_0) e^{j\omega t} d\omega \\
&= \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{2} G(\omega) e^{j\omega t} d\omega \cos(\omega_0 t) \\
&\quad + \frac{j}{2\pi} \int_{-\infty}^{\infty} \frac{1}{2} G(\omega) \operatorname{sign}(\omega) e^{j\omega t} d\omega \frac{e^{j\omega_0 t} - e^{-j\omega_0 t}}{2j}
\end{aligned}$$

$$= \frac{1}{2} g(t) \cos(\omega_0 t) - \frac{1}{2\pi} \int_{-\infty}^{\infty} \frac{1}{2} G(\omega) \frac{\text{sign}(\omega)}{j} e^{j\omega t} d\omega \sin(\omega_0 t) .$$

From this we obtain the SSB AM signal in terms of $g(t)$ and its sinusoidal modulating signals:

$$s_s(t) = \frac{1}{2} g(t) \cos(\omega_0 t) - \frac{1}{2} g^H(t) \sin(\omega_0 t) \quad (4.2.1.E6) .$$

Where

$$g^H(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} G(\omega) \frac{\text{sign}(\omega)}{j} e^{j\omega t} d\omega \quad (4.2.1.E7)$$

is called the Hilbert transform [6] of $g(t)$. It is easily shown from 4.2.1.E7 that the Hilbert transform is a linear transform and that if $g(t)$ is a real function then $g^H(t)$ is also real. Clearly it is a version of $g(t)$ in which the spectral components have been phase shifted over 90 degrees in a direction depending on the polarity of $\text{sign}(\omega)$. 4.2.1.E6 shows that this Hilbert transform of $g(t)$ is put in the quadrature channel by the SSB AM process. Obviously the removed lower sidebands must have had an opposite component in the quadrature channel, since as we have seen in the previous section, the complete DSB AM signal has no quadrature component.

An obvious way to do SSB-AM with analog circuitry is depicted in Fig.4.2.1.F2 [1][2][7]. An accurate way to do the Hilbert transform by digital means is to determine the impulse response of the filter that would create $g^H(t)$ if

the input is $g(t)$ and then to pass $g(t)$ through a filter that has coefficients corresponding to a properly sampled version of this impulse response. The latter is easily calculated since it would be the Hilbert transform of a $\delta(t)$ pulse (see Fig. 4.2.1.F3).

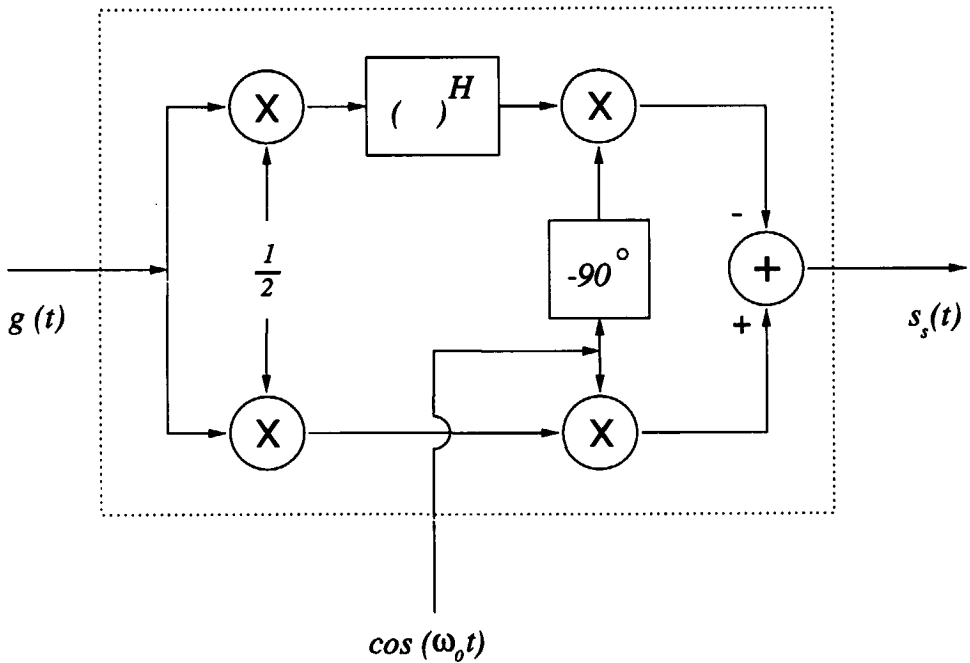


Figure 4.2.1.F2

An SSB-AM Modulator

In this calculation we limit, as always, the bandwidth of the channel to a bandwidth B which in NTSC M would be 4.2MHz and in other TV standards may vary from 4.2 to 6 MHz.

$$\delta_B^H(t) = \frac{1}{2\pi} \int_{-2\pi B}^{2\pi B} \frac{\text{sign}(\omega)}{j} e^{j\omega t} d\omega \quad (4.2.1.E8).$$

This equation is graphically depicted in Fig.4.2.1.F3.

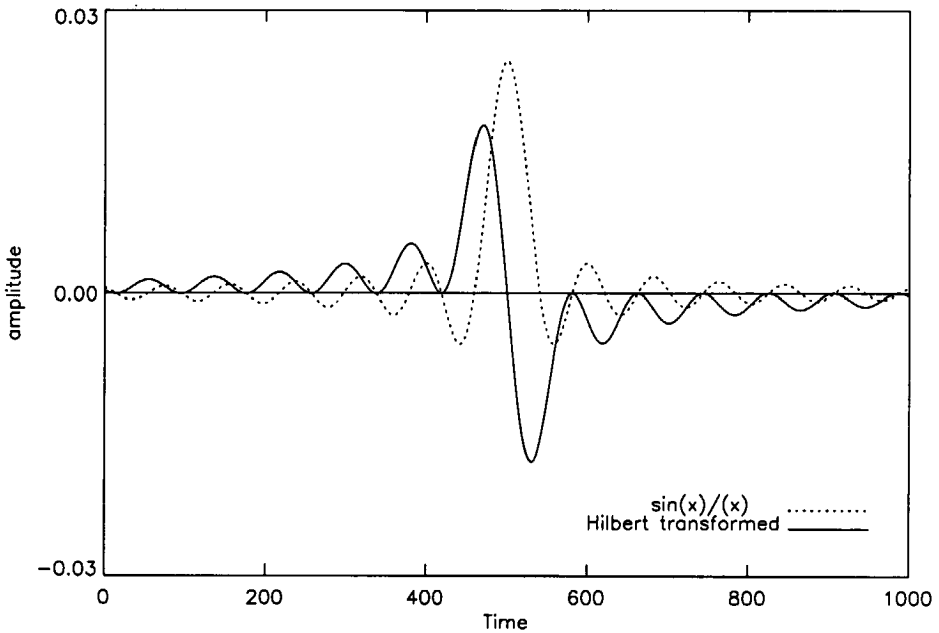


Figure 4.2.1.F3

The Hilbert Transform of a Band-limited Windowed $\frac{\sin(x)}{(x)}$ Signal

This function oscillates with a period $1/B$ between zero and $4/t$, is an odd function of t and passes through zero at $t = 0$. Elementary calculus shows that its

maximum deviation from zero occurs close to $t=1/2B$ and $-1/2B$ and has an absolute value close to $8B$.

In comparison the band limited $\delta(t)$ function itself (which is a $\frac{\sin(x)}{(x)}$ function as shown in Fig.4.2.1.F3), as is well known, oscillates with a period $1/B$ between $+2/t$ and $-2/t$, is an even function of t and has a maximum value at $t=0$ of $4\pi B$.

This shows that band limited Hilbert transformed $\delta(t)$ functions can qualitatively be described as lopsided versions of bandlimited $\delta(t)$ functions with roughly the same magnitude. Remembering this will be useful when we treat demodulated echoes in SSB AM.

4.2.2 SSB AM Demodulation

When no echoes are present the result of demodulation, again followed by low pass filtering, can be calculated by multiplying 4.2.1.E6 by $2\cos(\omega_0 t + \phi)$ taking into account a possible phase error ϕ and only retaining the terms with RF frequency zero. Since this calculation has no new complications compared with what we did before we simply mention the result:

$$\tilde{r}_s(t) = \frac{1}{2} g(t) \cos\phi - \frac{1}{2} g^H(t) \sin\phi \quad (4.2.2.E1).$$

In the full treatment of demodulation with echoes (see 4.2.4) this result can be found back by putting all echoes equal to zero.

4.2.2.E1 is not surprising: a phase error ϕ brings the contents of the quadrature channel back into the in-phase channel.

4.2.3 Echoes in the SSB AM Modulated Signal

As before we introduce a single echo by introducing a time delay in 4.2.1.E6 and weight it with an echo amplitude a :

$$e_s(t) = a s_s(t-\Delta t) \quad (4.2.3.E1)$$

$$\begin{aligned} &= a \frac{1}{2} g(t-\Delta t) \cos(\omega_0 t - \omega_0 \Delta t) - a \frac{1}{2} g^H(t-\Delta t) \sin(\omega_0 t - \omega_0 \Delta t) \\ &= [a \frac{1}{2} g(t-\Delta t) \cos(\omega_0 \Delta t) + a \frac{1}{2} g^H(t-\Delta t) \sin(\omega_0 \Delta t)] \cos(\omega_0 t) \\ &\quad + [a \frac{1}{2} g(t-\Delta t) \sin(\omega_0 \Delta t) - a \frac{1}{2} g^H(t-\Delta t) \cos(\omega_0 \Delta t)] \sin(\omega_0 t) . \end{aligned}$$

If there are more echoes we have to sum as in 4.1.3. The total received signal will then be obtained by adding the original SSB AM modulated signal 4.2.1.E6:

$$\hat{s}_s(t) = \frac{1}{2}g(t)\cos(\omega_0 t) - \frac{1}{2}g^H(t)\sin(\omega_0 t) \quad (4.2.3.E2)$$

$$+ \sum_{i=0}^{n-1} \left[a_i \frac{1}{2}g(t-\Delta t_i)\cos(\omega_0 \Delta t_i) + a_i \frac{1}{2}g^H(t-\Delta t_i)\sin(\omega_0 \Delta t_i) \right] \cos(\omega_0 t)$$

$$+ \sum_{i=0}^{n-1} \left[a_i \frac{1}{2}g(t-\Delta t_i)\sin(\omega_0 \Delta t_i) - a_i \frac{1}{2}g^H(t-\Delta t_i)\cos(\omega_0 \Delta t_i) \right] \sin(\omega_0 t) .$$

4.2.4 Echoes in SSB AM Demodulation

As in the DSB AM case we have to consider the question of finding the proper phase of the demodulating signal $\cos(\omega_0 t + \phi)$. This issue should be deferred for the moment because we are treating the case of SSB AM with suppressed carrier in anticipation of later results on VSB AM where there definitely is an unsuppressed carrier.

The general pattern of the computation will be similar to 4.1.4 where we treated the case of DSB AM. In particular we will again obtain (even more) complicated equations and again the complexity will vanish to a large extent when we apply low pass filtering in order to get rid of the double frequency components.

We start by multiplying the received RF signal 4.2.3.E2 by $2\cos(\omega_0 t + \phi)$ and obtain, apart from a multiplicative factor, the result:

$$r_s(t) = \hat{s}_s(t) 2\cos(\omega_0 t + \phi) \quad (4.2.4.E1)$$

$$\begin{aligned}
&= \frac{1}{2}g(t)\cos\phi + \frac{1}{2}g(t)\cos\phi\cos(2\omega_0t) - \frac{1}{2}g^H(t)\cos\phi\sin(2\omega_0t) \\
&+ \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\cos(\omega_0\Delta t_i) + a_i g^H(t-\Delta t_i)\sin(\omega_0\Delta t_i)]\cos\phi \\
&+ \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\cos(\omega_0\Delta t_i) + a_i g^H(t-\Delta t_i)\sin(\omega_0\Delta t_i)]\cos\phi\cos(2\omega_0t) \\
&+ \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\sin(\omega_0\Delta t_i) - a_i g^H(t-\Delta t_i)\cos(\omega_0\Delta t_i)]\cos\phi\sin(2\omega_0t) \\
&- \frac{1}{2}g(t)\sin\phi\sin(2\omega_0t) + \frac{1}{2}g(t)^H\sin\phi - \frac{1}{2}g^H(t)\sin\phi\cos(2\omega_0t) \\
&- \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\cos(\omega_0\Delta t_i) + a_i g^H(t-\Delta t_i)\sin(\omega_0\Delta t_i)]\sin\phi\sin(2\omega_0t) \\
&- \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\sin(\omega_0\Delta t_i) - a_i g^H(t-\Delta t_i)\cos(\omega_0\Delta t_i)]\sin\phi \\
&+ \frac{1}{2}\sum_{i=0}^{n-1} [a_i g(t-\Delta t_i)\sin(\omega_0\Delta t_i) - a_i g^H(t-\Delta t_i)\cos(\omega_0\Delta t_i)]\sin\phi\cos(2\omega_0t) .
\end{aligned}$$

Removing the double frequency terms then yields:

$$\tilde{r}_s(t) = \frac{1}{2}g(t)\cos\phi \quad (4.2.4.E2)$$

$$\begin{aligned}
& + \frac{1}{2} \sum_{i=0}^{n-1} [a_i g(t-\Delta t_i) \cos(\omega_0 \Delta t_i) + a_i g^H(t-\Delta t_i) \sin(\omega_0 \Delta t_i)] \cos\phi \\
& + \frac{1}{2} g^H(t) \sin\phi \\
& - \frac{1}{2} \sum_{i=0}^{n-1} [a_i g(t-\Delta t_i) \sin(\omega_0 \Delta t_i) - a_i g^H(t-\Delta t_i) \cos(\omega_0 \Delta t_i)] \sin\phi .
\end{aligned}$$

From the previous considerations it will be clear that in 4.2.4.E2:

- $\cos\phi$ means received in the in-phase channel and left there by the demodulation;
- $\sin\phi$ means received in the quadrature channel and flipped to the main channel by demodulation;
- $\cos(\omega_0 \Delta t_i)$ means echoed within either the in-phase or the quadrature channel;
- $\sin(\omega_0 \Delta t_i)$ means echoed from the in-phase channel to the quadrature channel or vice versa;
- $g(t)$ means modulated into the in-phase channel
- $g^H(t)$ means modulated into the quadrature channel.

Consequently the terms in 4.2.4.E2 take on the following meaning:

1 First term:

Modulated into the in-phase channel and left there by direct transmission and demodulation

2 First term in first sum:

Modulated into the in-phase channel and left there by the echo process and demodulation

3 Second term in first sum:

Modulated into the quadrature channel, echoed back into the in-phase channel and left there by demodulation

4 Next term:

Modulated into the quadrature channel, left there by direct transmission and flipped into the in-phase channel by demodulation

5 First term in the second sum:

Modulated into the in-phase channel, echoed into the quadrature channel and flipped back into the in-phase channel by demodulation

6 Second term in second sum:

Modulated into the quadrature channel, left there by the echoing process and flipped to the in-phase channel by demodulation

The occurrence of these six terms indeed covers all relevant possibilities as the tree structure of Fig.4.2.4.F1 shows.

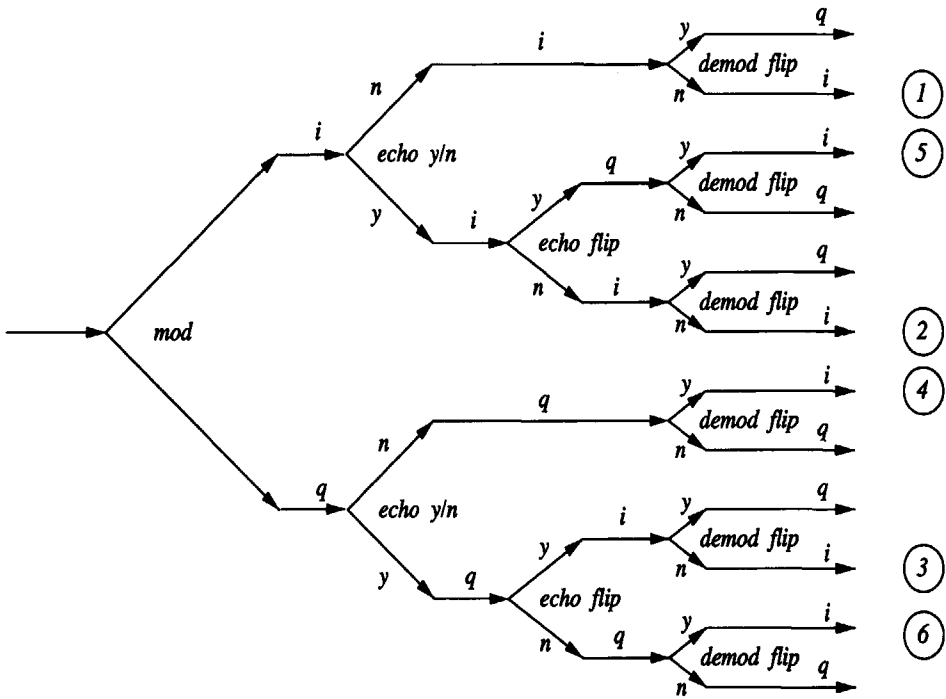


Figure 4.2.4.F1

The Tree Structure Representation of the Echoed Components

At the modulation stage there is a binary split according to whether the signal goes to the in-phase or quadrature channel. Then there is a split without the possibility of a flip from one channel to the other according to whether the signal is going to be echoed or not. This leaves 4 possibilities. Only two of these four are subject to a flip by echoing, which leaves six possibilities. All of these are to be flipped or not in the demodulator, leading to twelve possibilities. Only half of these end up in the in-phase channel. These are the six mentioned above. The

tree branches ending up in the in-phase channel have been numbered in accordance with the order in which they were mentioned in the above.

The main conclusion from all this is that the base-band response is linear in $g(t)$ again.

4.3 Use of an Intermediate Frequency (IF) in Demodulation

In the above we have consistently assumed that demodulation is done directly from RF to baseband. In practice this is not the case in general. Instead the RF signal is first transferred to an intermediate frequency (IF) which is suitable to do the necessary filtering but which is the same for all the different TV channels (i.e. radio frequencies) which one might wish to receive. We therefore still have to ascertain that this procedure does not introduce unexpected complications in the present context. This can easily be done as follows.

Assume that we have an incoming RF signal

$$s(t) = A(t)\cos(\omega_0 t) + B(t)\sin(\omega_0 t) \quad (4.3.E1).$$

This assumption covers the DSB AM (B equal to 0) as well as the SSB AM (B unequal to 0) cases. A transfer to an IF ω_1 can be effected by multiplication with $2\cos[(\omega_1 + \omega_0)t + \phi]$, the result being:

$$r(t) = [A(t)\cos(\omega_0 t) + B(t)\sin(\omega_0 t)] 2\cos[(\omega_1 + \omega_0)t + \phi] \quad (4.3.E2)$$

$$= A(t)\cos(\omega_1 t + \phi) + A(t)\cos[(2\omega_0 + \omega_1)t + \phi]$$

$$+ B(t)\sin(\omega_1 t + \phi) + B(t)\sin[(2\omega_0 + \omega_1)t + \phi] .$$

Removal of the high frequency components leads to

$$\bar{r}(t) = A(t)\cos(\omega_1 t + \phi) + B(t)\sin(\omega_1 t + \phi) \quad (4.3.E3).$$

Since in all of the above we never assumed any particular temporal relationship between the modulating signals and the carrier waves we can safely and without loss of generality adjust the time axis for the intermediate frequency and drop ϕ . That leads to an expression that is identical to 4.3.E1 apart from a change from ω_0 to ω_1 , which leaves all of the previous reasoning valid.

We thus have shown that the use of \bar{r} does not interfere with our analyses.

4.4 Echoes in Vestigial-side-band Amplitude (VSB-AM) Modulation and Demodulation

4.4.1 VSB-AM Modulation

VSB-AM [1][2] provides an extremely useful balance between bandwidth efficiency and communication reliability and is widely used in analog and digital communications as well as in broadcasting. In the present context we need to do a detailed analysis of the echo phenomenon in VSB-AM in view of the fact that, as we shall see, VSB is a mix between DSB-AM and SSB-AM that is further complicated by the fact that VSB-AM needs a special filter [8] at the receiver side in order to compensate for an essential distortion that occurs in the VSB-AM demodulation process. We can thus expect to meet (at least) all the complications that we hitherto encountered, especially those connected to the quadrature channel in the case of SSB-AM.

In the following we will use a notation for the modulating signal that is closely related to what we did in 4.1.1. However we will omit the multiplicative constant D as used in 4.1.2.E1 and thereafter and we will assume that for the SSB part of VSB we will have the same multiplicative factor m as the one used for the DSB part, which was called the modulation index in 4.1.1

Also we will continue to be consistent in describing modulation as multiplication with $\cos(\omega_0 t)$ and demodulation with $2\cos(\omega_0 t + \phi)$. This will enable us to keep track of the relative amplitudes of the DSB parts as compared to those of the SSB parts after demodulation.

In Figure 4.4.1.F1 we show once more the DSB-AM spectrum, in this case

as a dotted curve. The solid curve represents VSB-AM, in which part of the lower side-band of the DSB spectrum has been filtered out, the truncated lower side-band now extending from $\omega_0 - \omega_v$ to ω_0 . In many VSB-AM systems, including television, there is an unsuppressed carrier at frequency ω_0 . This carrier is represented in the equations as before by the term 1 in the equation $g(t) = 1 + mf(t)$.

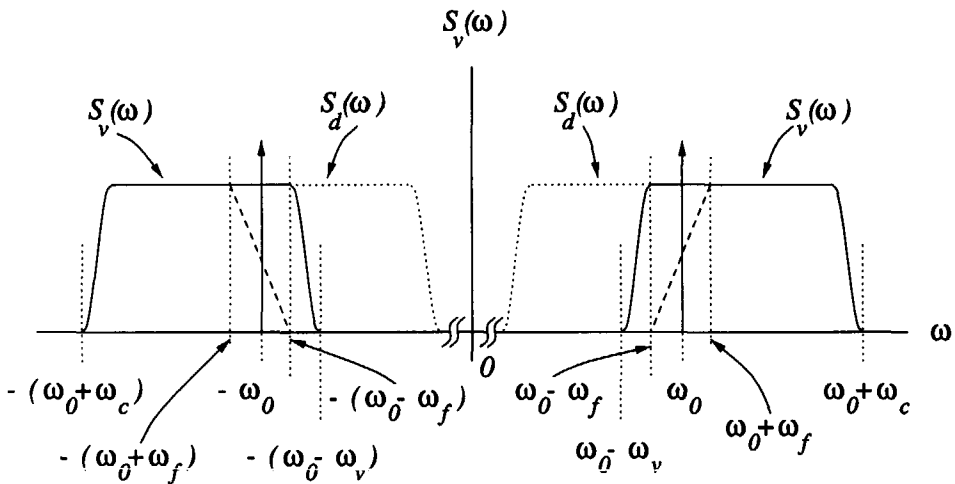


Figure 4.4.1.F1

The Spectrum of a VSB-AM Modulated Signal with Nyquist Filter Curve
Compared with a DSB-AM Spectrum

As will be clear from Figure 4.4.1.F1 there is now a DSB component in the modulated signal as well as a SSB component. By comparing 4.1.4.E4 for the case of no echoes with 4.2.2.E1 we see that after demodulation the DSB part of the signal would have picked up a multiplicative factor of 2 as compared with the SSB part. The trick in VSB to prevent this is to incorporate in the receiver a so-

called Nyquist [8] filter (or VSB filter) with the characteristic depicted in Figure 4.4.1.F1 as the dashed curve rising from 0% at $\omega_0 - \omega_f$ to 100% at $\omega_0 + \omega_f$, where $\omega_f < \omega_v$.

The effect of this filter is that the modulated signal now has a DSB part from $\omega_0 - \omega_f$ to $\omega_0 + \omega_f$ consisting of the vertically shaded triangles in Figure 4.4.1.F2 and a SSB part from ω_0 to ω_c consisting of the horizontally shaded areas in Figure 4.4.1.F2. This is exactly the same as we would have had if, referring to Figure

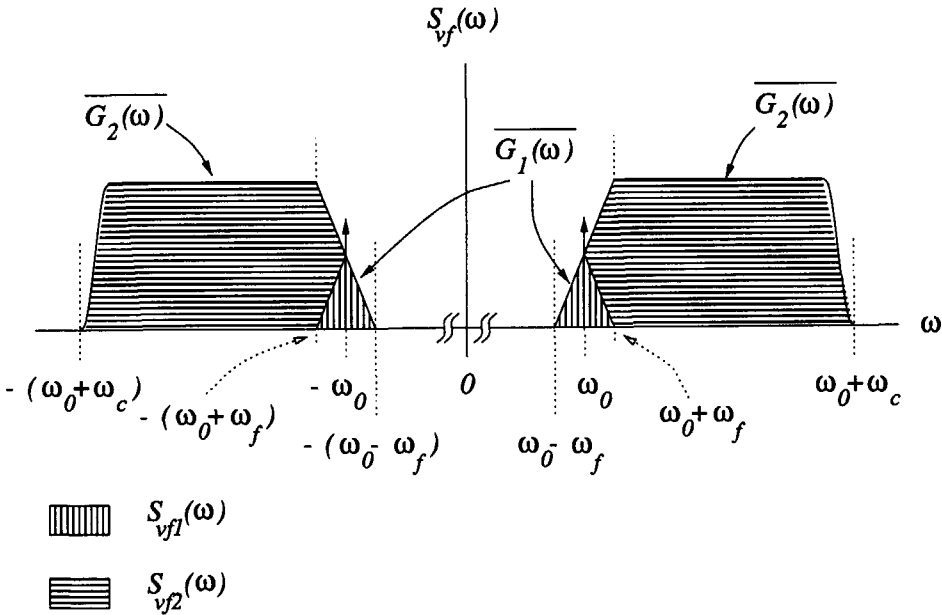


Figure 4.4.1.F2

The Spectrum of a Nyquist Filtered VSB-AM Signal

4.4.1.F3, we had started out with a baseband signal $\overline{G_1(\omega)}$ being modulated in DSB fashion and a baseband signal $\overline{G_2(\omega)}$ being modulated in SSB fashion. So

after demodulation we would have, $\overline{2G_1(\omega)} + \overline{G_2(\omega)}$, taking into account the factor 2 that is picked up as described above by the DSB part as compared to the SSB part.

Obviously this result is exactly identical to the original baseband spectrum $G(\omega)$. We thus see that in VSB-AM in the absence of echoes no distortion of the original signal results. This of course is an already well known result. What we have gained is that on the basis of what was hitherto said we can immediately write down the proper expression for the echoed signal in the case of VSB-AM on the basis of our results obtained for DSB and SSB in previous sections.

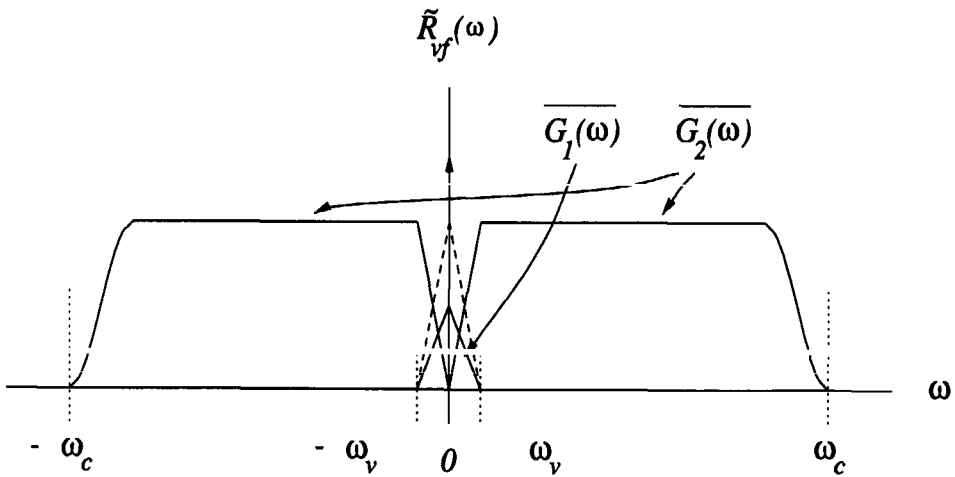


Figure 4.4.1.F3

The Spectrum Relation between $\overline{G_1(\omega)}$ and $\overline{G_2(\omega)}$

4.4.2 Echoes in VSB-AM

The recipe to be followed in order to see what happens in the case of VSB-AM now clearly is as follows:

1. Split $G(\omega)$ into $\overline{G_1(\omega)}$ and $\overline{G_2(\omega)}$ as shown in Figure 4.4.1.F3 so that $G(\omega) = 2\overline{G_1(\omega)} + \overline{G_2(\omega)}$.
2. Write down the expression for the demodulated received echoed signal for $\overline{g_1(t)}$ in the DSB case according to 4.1.4.E4 with an additional multiplicative factor of 2.
3. Write down the expression for the demodulated received echoed signal for $\overline{g_2(t)}$ in the SSB case according to 4.2.4.E2 without an additional multiplicative factor.
4. Add the results of steps 2 and 3.

The result is

$$\tilde{r}_v(t) = \overline{g_1(t)} \cos\phi + \frac{1}{2} \overline{g_2(t)} \cos\phi \quad (4.4.2.E1)$$

$$+ \sum_{i=0}^{n-1} [a_i \overline{g_1(t-\Delta t_i)} \cos(\omega_0 \Delta t_i) + a_i \frac{1}{2} \overline{g_2(t-\Delta t_i)} \cos(\omega_0 \Delta t_i)] \cos\phi$$

$$\begin{aligned}
& + \sum_{i=0}^{n-1} a_i \frac{1}{2} \overline{(g_2(t-\Delta t_i))^H} \sin(\omega_0 \Delta t_i) \cos \phi \\
& + \frac{1}{2} \overline{(g_2(t))^H} \sin \phi \\
& - \sum_{i=0}^{n-1} [a_i \overline{g_1(t-\Delta t_i)} \sin(\omega_0 \Delta t_i) + a_i \frac{1}{2} \overline{(g_2(t-\Delta t_i))} \sin(\omega_0 \Delta t_i)] \sin \phi \\
& + \sum_{i=0}^{n-1} a_i \frac{1}{2} \overline{(g_2(t-\Delta t_i))^H} \cos(\omega_0 \Delta t_i) \sin \phi.
\end{aligned}$$

We are thus allowed to conclude that even in this complicated case of a mixture of DSB and SSB modulation as well as a mixture of events in the in-phase and in the quadrature channel the total process of echo creation from base-band level to base-band level is linear. Of course one of the facts that allow us to make that statement is the linearity of the Hilbert transform.

This will permit the process of echo cancellation, to be treated in later chapters, to be set up on the basis of linear programmable filters.

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CHAPTER 5

REQUIREMENTS FOR GHOST CANCELLATION REFERENCE SIGNALS (GCRs)

As remarked already in chapter 2 a practical way to achieve echo cancellation is to send as part of the broadcast a standardized "reference" signal that is also stored in the memory of the receiver. This enables the receiver to compare the stored, clean reference signal with the received, distorted version in order to obtain the information that is necessary for proper echo cancellation. Since in analog TV echoes give rise to repeats of the picture on the screen which are known as "ghosts", the reference signal is frequently called "Ghost Cancellation Reference" or GCR. In the following we will use this name as well.

Obviously the GCR has to be specifically designed in order to optimize it in terms of various criteria. This chapter starts an account, to be continued in the next chapter, of our design of a family of GCRs that in the mean time has been adopted as the basis of the ITU world standard for GCRs.

The basic considerations on which our thinking was based were of three different types:

1. The GCR should be such that all needed information on the complete

transmission channel is provided.

2. The GCR should be as robust as possible to disturbances other than echoes such as additive Gaussian noise, quantization noise introduced by digital processing in the receiver and jitter in the clock system that serves to provide the timing of this digital processing.
3. The GCR should be suitable for the use of as many known algorithms for extracting the channel model as possible.

From these principles we arrived at list of requirements on GCRs, to be treated in the following sections, that were sufficiently strict to allow the straightforward design of our family of GCRs as described in the next chapter

5.1 Requirements Suggested by the Frequency Division Algorithm

If $g(t)$ and $G(\omega)$ are the original GCR and its Fourier transform and $r(t)$ and $R(\omega)$ are the received, distorted GCR and its Fourier transform, while $h(t)$ and $H(\omega)$ are the impulse response and frequency response functions of the channel, then

$$R(\omega) = G(\omega) H(\omega) + N(\omega) \quad (5.1.E1)$$

where $N(\omega)$ is the noise added in the channel.

The frequency division algorithm for echo cancellation is based on 5.1.E1 divided by $G(\omega)$:

$$\frac{R(\omega)}{G(\omega)} = H(\omega) + \frac{N(\omega)}{G(\omega)} \quad (5.1.E2)$$

In this equation we see that $H(\omega)$ can be obtained from the quotient of $R(\omega)$ and $G(\omega)$ [1] with an accuracy that depends on the relative magnitude of $H(\omega)$ and $\frac{N(\omega)}{G(\omega)}$. Since $H(\omega)$ cannot be controlled in advance we have to minimize $\frac{N(\omega)}{G(\omega)}$ at all frequencies in the band of interest and since $N(\omega)$ cannot be controlled in advance this means that we have to maximize $|G(\omega)|$ at all frequencies in the band of interest to the extent to which regulations, dynamic range of equipment and cost allow this. What we have learned here is the not very surprising result that the magnitude of $G(\omega)$ should be as large as possible, but that from the point of view of signal to noise ratio it is not very critical that $G(\omega)$ is strictly independent of ω . However if we make the power spectrum of $G(\omega)$ absolutely constant over the entire (limited) bandwidth of the channel we have the additional benefit that we can obtain $h(t)$ directly from the cross-correlation of $g(t)$ and $r(t)$ as we will show in the next section.

5.2 Considerations on the Cross-correlation of $g(t)$ and $r(t)$

The cross-correlation of $g(t)$ and $r(t)$ is

$$\phi_{12}(\tau) = \int_{-\infty}^{\infty} g(t-\tau) r(t) dt$$

Because of our earlier result, obtained in chapter 4, that the echo creation process allows for a linear base-band model, even in the complex case of vestigial side-band modulation, we consider $g(t)$ and $r(t)$ at base-band level, i.e. before modulation and after demodulation respectively. The bandwidth of $G(\omega)$ will be restricted to the bandwidth of the channel. Consequently the bandwidth of $R(\omega)$ will be similarly restricted since a linear process does not introduce new frequencies.

If we now write $g(t)$ and $r(t)$ in terms of their Fourier transforms we obtain

$$\phi_{12}(\tau) = \frac{1}{4\pi^2} \int_{-\infty}^{\infty} dt \int_{-2\pi B}^{2\pi B} d\bar{\omega} \int_{-2\pi B}^{2\pi B} d\bar{\omega} G(\bar{\omega}) R(\bar{\omega}) e^{j\bar{\omega}(t-\tau)} e^{j\bar{\omega}t}$$

where B is the bandwidth of the channel.

Carrying out the integration over t yields

$$\phi_{12}(\tau) = \frac{1}{2\pi} \int_{-2\pi B}^{2\pi B} d\bar{\omega} \int_{-2\pi B}^{2\pi B} d\bar{\omega} G(\bar{\omega}) R(\bar{\omega}) e^{-j\bar{\omega}\tau} \delta(\bar{\omega} + \bar{\omega})$$

which equals

$$\phi_{12}(\tau) = \frac{1}{2\pi} \int_{-2\pi B}^{2\pi B} d\bar{\omega} G(\bar{\omega}) R(-\bar{\omega}) e^{-j\bar{\omega}\tau}$$

Writing $G(-\omega) H(-\omega)$ for $R(-\omega)$ yields

$$\begin{aligned} \phi_{12}(\tau) &= \frac{1}{2\pi} \int_{-2\pi B}^{2\pi B} d\bar{\omega} G(\bar{\omega}) G(-\bar{\omega}) H(-\bar{\omega}) e^{-j\bar{\omega}\tau} \\ &= \frac{1}{2\pi} \int_{-2\pi B}^{2\pi B} d\bar{\omega} \left| G(\bar{\omega}) \right|^2 H(\bar{\omega}) e^{j\bar{\omega}\tau} \end{aligned} \quad (5.2.E1)$$

since $G(\omega) G(-\omega)$ equals $G(\omega) G^*(\omega)$ if $g(t)$ has been designed to be a real signal which, as argued in Chapter 1, is desirable for economic reasons.

If the power spectrum is flat over the bandwidth of the channel $G(\omega) G^*(\omega)$ can be taken out of the integration as a constant K and we then are left with

$$\phi_{12}(\tau) = K h(\tau)$$

This shows that, as promised at the end of 5.1, indeed in case of a flat spectral distribution of $G(\omega)$ the cross-correlation of $g(t)$ and $r(t)$ equals the impulse response function of the channel if $g(t)$ is a real signal.

It should be noted that in that case the auto-correlation function of $g(t)$ is obtained by putting in 5.2.E1

$$H(\bar{\omega}) = 1 \quad - 2 \pi B < \omega < 2 \pi B .$$

Clearly then this auto-correlation becomes a $\frac{\sin(x)}{(x)}$ function.

5.3 Completion of the List of Requirements for the GCR

The requirements that we formulated up to this stage are:

1. High signal energy at all frequencies
2. A flat spectral distribution over the entire bandwidth of the base-band channel if one wishes to use correlation to determine the channel characteristic.
3. The GCR should be a real signal i.e. $G(\omega) = G^*(-\omega)$.

To this list we have to add the following:

4. The time duration of the GCR should be as short as possible with an even temporal distribution of energy.

Time is a resource that has to be shared between the TV signal itself and the GCR. Obviously as little time as possible has to be spent on the GCR. This implies that during the time intervals (in the vertical blanking periods) that are devoted to the GCR there should be as little time as possible wasted on episodes in the GCR that are poor in energy content. Consequently the energy in the GCR should be distributed over its duration as evenly as possible.

Note that by requirements 1 and 4 we have demanded an even distribution of energy in the frequency as well as in the time domain.

The requirement of a high energy for all the frequencies in the bandwidth combined with the restriction of the duration of the GCR to about 30 usec as described in 2.3.1 implies the requirement of a smooth phase characteristic for the GCR. In the next section we will discuss and heuristically prove this in more detail.

5. The design of the GCR should be such that dynamic range, length and bandwidth are adjustable parameters in order to have easy accommodation of the signal to the various TV systems that are used in the world, i.e. the design should cover a whole family of signals that all satisfy the requirements 1 to 4.

6. There should be no severe limitations on the sampling interval in the time domain as well as in the frequency domain that is used in the receiver to perform the signal processing necessary for echo cancellation, so that in the choice of the sampling intervals in both time and frequency domains there is opportunity left to comply with the local TV system and the specifications of the receiver. This suggests that the most convenient

choice would be an analog GCR.

We would like to mention here that we have been able to establish with certainty that it is impossible to use as GCR any digital pulse sequence that would satisfy all of the above requirements. We are fully aware of a number of proposals that have been made in this direction but satisfactory proof that requirements 1 through 6 are all satisfied for any one of these proposals has never reached us. On the other hand the analog solution that we will present below will be seen to satisfy all the requirements. In chapter 7 we will return at length to this issue.

5.4 Explanation of the Issue of a Smooth Phase Characteristics

In order to see that the simultaneous requirements on the GCR of high energy at all frequencies within the bandwidth and limited time duration imply the requirement of a smooth phase characteristic we have to use the concept of group delay [6], which we recall here for the sake of clarity.

If we have a signal

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F(\omega) e^{j\omega t} d\omega ,$$

and if we write

$$F(\omega) = |F(\omega)| e^{j\phi(\omega)}$$

where obviously $\phi(\omega)$ is the phase of the frequency component $F(\omega)$ we obtain

$$f(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} |F(\omega)| e^{j[\phi(\omega) + \omega t]} d\omega .$$

The contribution to this integral of each frequency interval $d\omega$ is determined by the extent to which the exponential rotates in the complex plane when ω passes through that interval. If there is much rotation, the contribution of the interval to the integral tends to cancel. Clearly the contribution is significant only if there is no rotation i.e. when $\frac{\partial [\phi(\omega) + \omega t]}{\partial \omega} = 0$ or $\frac{\partial \phi(\omega)}{\partial \omega} = -t$. Thus frequency components at frequency ω will tend to be concentrated at a point in time where $t = -\frac{\partial}{\partial \omega} \phi(\omega)$. Consequently if ϕ is constant as a function of ω all frequencies are concentrated around $t = 0$; if ϕ is linearly dependent on ω , i.e. if $\phi(\omega) = \phi_0 + \omega t_0$; all frequencies will be concentrated around t_0 ; and if ϕ depends in a more complicated way on ω , various frequencies will be concentrated in different points in time. For this reason the value of $-\frac{\partial \phi(\omega)}{\partial \omega}$ at a frequency ω is called the group delay for that frequency.

Now, coming back to the issue mentioned above of combined requirements of high energy at all frequencies and a limited duration of the GCR, it will be clear that a sudden phase jump as a function of frequency can cause the frequencies where the phase jump occurs to be dislocated in time to such an extent that subsequent time windowing in order to limit the duration of the signal can cause these frequencies to disappear from the signal. This would then violate the requirement

of high energy at all frequencies. We illustrate this now in a more quantitative way.

Assume that the phase of the GCR varies linearly with frequency by an amount $a\pi$ over a frequency interval $\Delta\omega$ while being constant at all other frequencies. We would then have a derivative of phase with respect to frequency in the frequency interval $\Delta\omega$ with the result that the pertaining frequencies would have a group delay of

$$T = - \frac{a\pi}{\Delta\omega}$$

relative to the other frequency components. Clearly if this delay exceeds half the required duration of the GCR, which itself would be about 30 μsec , then truncation of the resulting signal to 30 μsec would result in a substantial loss of energy in the frequency interval $\Delta\omega$.

We thus would have the condition that

$$\frac{a\pi}{\Delta\omega} < 15 \mu\text{sec}$$

In terms of normal rather than angular frequencies this would mean

$$\Delta f > \frac{a}{30} \text{ Mhz}$$

In words this means that a linear phase change of $a\pi$ should be stretched over a frequency interval of at least $\frac{a}{30}$ Mhz to prevent severe loss of energy in that frequency interval from a signal that is restricted in time to 30 μ sec. The extension of this result to other signal durations should be obvious:

$$\Delta f > \frac{a}{\text{required signal duration}} .$$

The reader is reminded that the above reasoning applies to a single phase jump from one constant value to another, showing that if it is too sudden in terms of frequency we lose energy from the frequency range where it occurs. As will be seen below we avoided this potential problem by actually designing the phase behavior of our family of GCRs over the entire bandwidth in such a way that uniform spectral energy is guaranteed as well as uniform distribution of energy over signal duration.

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† RFP: Request For Proposal.

CHAPTER 6

THE DESIGN OF A FAMILY OF GCRS

As announced already in the previous chapter we will now describe how the requirements formulated there lead in a straightforward way to a complete design of a family of GCRs, defined by a set of adjustable parameters that allow specific adjustment to the various TV systems that are in use in the world.

6.1 Frequency Domain Considerations

Writing $f(t)$ and $F(\omega)$ respectively for the GCR in the time and in the frequency domain we have in general

$$f(t) = \frac{1}{2\pi} \int_{-\Omega}^{\Omega} F(\omega) e^{j\omega t} d\omega ,$$

where $\Omega = 2\pi$ times the channel bandwidth.

If $F(\omega)$ has absolute value $A(\omega)$ and argument $\phi(\omega)$ this becomes

$$f(t) = \frac{1}{2\pi} \int_{-\Omega}^{\Omega} A(\omega) e^{j\phi(\omega)} e^{j\omega t} d\omega \quad (6.1.E1)$$

Since we would like to have a uniform spectral density $A(\omega)$ must be a constant and since $f(t)$ must be real $\phi(\omega)$ must be an odd function of frequency.

As we saw during our considerations on group delay in the last chapter the actual location of the various frequency components in the time domain will be determined by $\phi(\omega)$. What we want to achieve in order to utilize the available time span of about 30 μsec as well as possible is the avoidance of time intervals inside that time span that lack energy. In other words we want to achieve an even distribution of energy over time. This implies that because of the fact that we already have an even distribution of energy in the frequency domain it is desirable to have the times where the various frequencies tend to be concentrated on the basis of the group delay concept distributed uniformly over time. Thus we should have a group delay that is proportional to ω , which suggests:

$$\frac{d\phi(\omega)}{d\omega} = 2b\omega,$$

leading to

$$\phi(\omega) = b\omega^2 \quad (6.1.E2).$$

There are two complications here, the first being that this would mean group delays of opposite sign for frequencies of opposite sign and the second being that now $\phi(\omega)$ is an even instead of an odd function of ω .

Both complications can be simultaneously lifted by having different signs in front of b for negative and positive frequencies. This necessitates splitting the integral 6.1.E1, where we already made $A(\omega)$ constant, into two parts and introducing 6.1.E2, so that 6.1.E1 becomes:

$$f(t) = \frac{1}{2\pi} \int_{-\Omega}^0 A [\cos(b \omega^2) - j \sin(b \omega^2)] e^{j \omega t} d\omega + \\ + \frac{1}{2\pi} \int_0^{\Omega} A [\cos(b \omega^2) + j \sin(b \omega^2)] e^{j \omega t} d\omega \quad (6.1.E3).$$

Before we turn to time domain considerations it is useful to realize that thus far we have obtained a real signal with a flat spectrum and an even distribution of frequencies over time as far as the center of each frequency interval on the time axis is concerned. We also have bandwidth, amplitude and duration as adjustable parameters. On the other hand we have not yet estimated or influenced the tails on the time axis that result from the "brick wall" frequency cut-off at the channel bandwidth.

Another point where we have to be cautious is that in our reasoning we have been on the fuzzy side in talking about energy and amplitudes without considering interferences of different frequencies. The only feasible defense here is that in the

end the results of our design had to be tested in practice. This turned out to be a success as we will discuss later.

6.2 Time Domain Considerations

In order to shape the signal in the time domain by removing the tails that come about because of the sharp cut-off in the frequency domain at the edge Ω_1 of the frequency band we now have to introduce in the integrands of 6.2.E1 a suitable frequency roll off by means of a windowing factor $W(\omega)$ that has such a roll off. Because $W(\omega)$ is to be a multiplicative factor the bandlimited character of $f(t)$ will not be impaired.

Equation 6.1.E3 now becomes

$$\begin{aligned}
 f(t) = & \frac{1}{2\pi} \int_{-\Omega}^0 A [\cos(b \omega^2) - j \sin(b \omega^2)] e^{j \omega t} W(\omega) d\omega + \\
 & + \frac{1}{2\pi} \int_0^{\Omega} A [\cos(b \omega^2) + j \sin(b \omega^2)] e^{j \omega t} W(\omega) d\omega
 \end{aligned}
 \tag{6.2.E1}$$

For $W(\omega)$ we take a function that in the frequency domain is the convolution of another brick wall function of frequency and a function that has a smooth bell shaped form in terms of frequency. In the time domain that would mean the product of a bandlimited delta pulse and a function which has a Fourier transform that

is bell shaped. For the latter we choose the so called Hanning [1] window:

$$\frac{1}{2} [1 + \cos(ct)] \quad - \frac{\pi}{c} < t < \frac{\pi}{c} ,$$

since its Fourier transform consists of three functions of $\frac{\sin(x)}{(x)}$ character with zero crossings at angular frequencies $\omega = nc$ for integer values of n , one centered at $\omega = 0$ and the others, with half the amplitude of the first at $\omega = c$ and $\omega = -c$. The latter cancel the tails of the first considerably thus producing the bell shape. This is illustrated in Figure 6.2.F1.

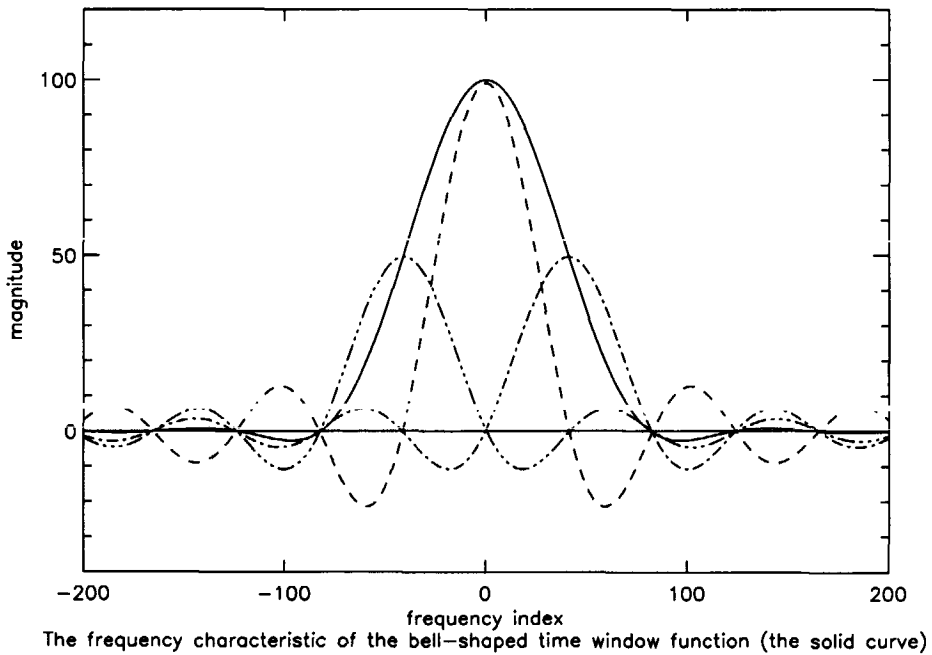


Figure 6.2.F1

We thus arrive at

$$w(t) = \frac{1}{2\pi} \int_{-\Omega_1}^{\Omega_1} e^{j\gamma t} d\gamma \left[\frac{1}{2} + \frac{1}{2} \cos(ct) \right] \quad -\frac{\pi}{c} < t < \frac{\pi}{c} \quad (6.2.E2).$$

Here Ω_1 and c are adjustable parameters

With this choice $w(t)$ and $w(-t)$ are complex conjugate and therefore the Fourier transform $W(\omega)$ of $w(t)$ is real so that its inclusion in 6.2.E1 does not affect the phase of the integrand, leaving the group delays that were designed in, unchanged.

In fact we have:

$$W(\omega) = \int_{-\frac{\pi}{c}}^{\frac{\pi}{c}} \left\{ \frac{1}{2\pi} \int_{-\Omega_1}^{\Omega_1} e^{j\gamma t} d\gamma \left[\frac{1}{2} + \frac{1}{2} \cos(ct) \right] \right\} e^{-j\omega t} dt \quad (6.2.E3).$$

With the GCR now obtained we can adjust the bandwidth to fit the various Television systems in the world, we can adjust the length of the signal by b once the bandwidth is fixed and then adjust the amplitude within the local regulatory situations in order to ensure maximal energy content. Here we would like to point out that if b is increased the frequencies are spread over a longer time period while the total energy content as independently determined by A remains the same. Consequently the amplitude decreases with increasing b .

Thus the GCR seems to meet all the requirements that we imposed, at least if as stated before, actual simulation and real life verification establish that the fuzzy points in our reasoning are valid. This was done in numerous computer checks and field tests for many of the TV systems that are in use today with complete success as to the expected character of the GCRs as well as to their potential for echo cancellation. It led ultimately to the adoption as a world standard as we already mentioned.

To conclude this chapter we illustrate a particular member GCR of the family in the time domain as well as in the frequency domain in 6.2.F2 and 6.2.F3 respectively.

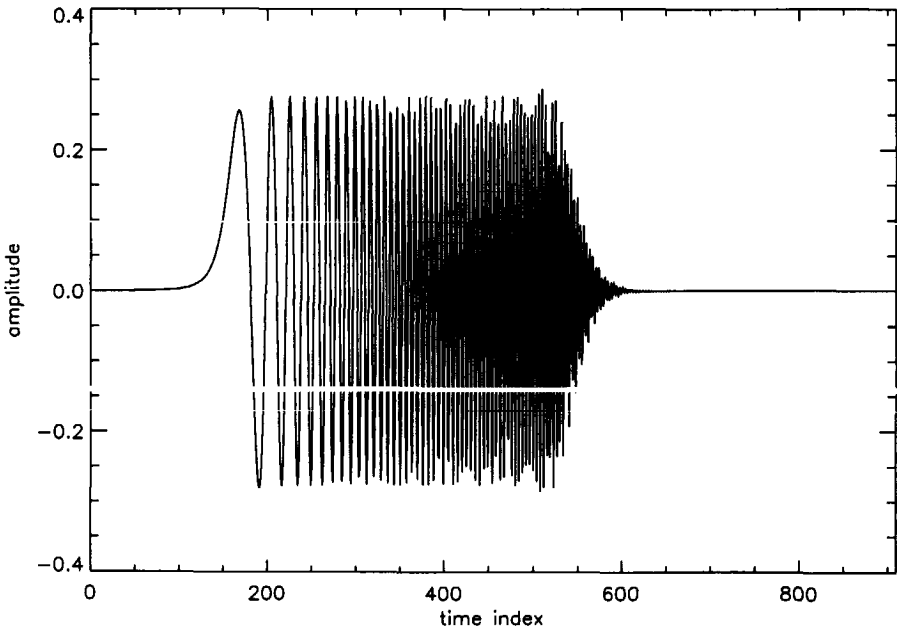
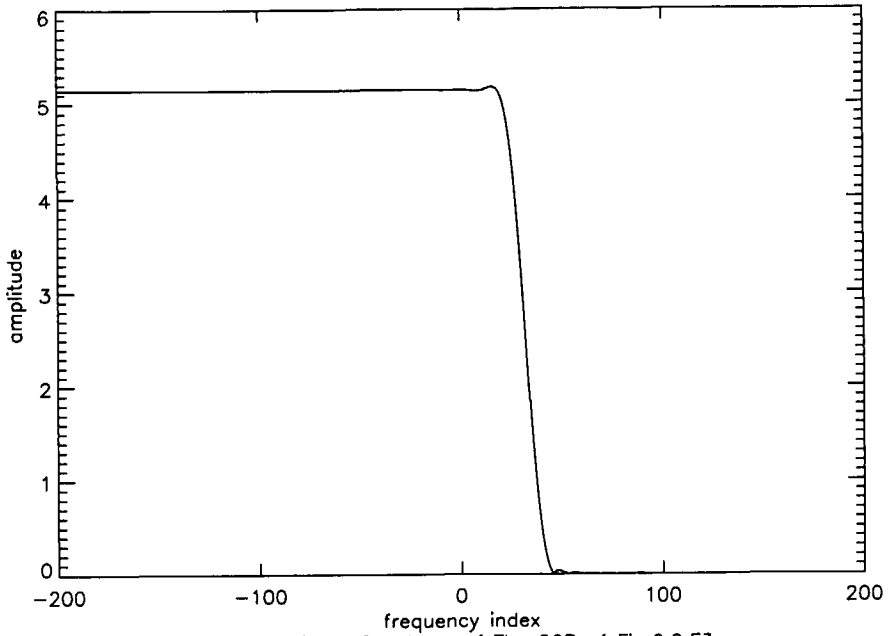


Figure 6.2.F2



The Amplitude Spectrum of The GCR of Fig.6.2.F3

Figure 6.2.F3

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CHAPTER 7

CONSIDERATIONS ON THE USE OF DIGITAL SEQUENCES AS GCRs

Pseudo random digital sequences (PRSs) have been proposed as GCRs by several people [1], [2], [3], [4], because of the expectation that they would have a flat spectrum. In this chapter we will show that theoretical considerations as well as computer simulations indicate convincingly that these expectations cannot be substantiated in a satisfactory way and that therefore PRSs are not optimal for use as GCRs.

7.1 General Characteristics of Pseudo Random Sequences

Pseudo random sequences will be defined for the purpose of our discussions in this chapter as randomly organized sequences of N delta pulses, occurring at fixed time intervals T , where the time integral of the pulses only can have either of two different values inside a sequence. We will for the time being assume that the sequences repeat indefinitely, which implies that they have a discrete spectrum which contains only frequencies that are a multiple of the inverse of the repetition period NT . The repetition property also implies that the auto-correlation function is periodic with a period that equals the length of a single sequence in time, NT .

The "random" character (in fact the PRS is a deterministic sequence) of the sequences of course is expressed by the fact that the auto-correlation function $\phi(\tau)$ differs from zero only for $\tau = nNT$ (where $n = \dots -1, 0, 1, \dots$ and where τ is a continuous parameter). In other words $\phi(\tau)$ is a series of delta pulses with a period NT . Consequently its spectrum is formed by sampling of the spectrum of a single delta pulse at frequencies that are a multiple of $\frac{1}{NT}$, i.e. the discrete spectrum of the auto-correlation function consists of mutually equal components at frequencies that are a multiple of $\frac{1}{NT}$.

Now it is a well known fact [5] that the spectrum of an auto-correlation function is identical with the power spectrum of the original function. We thus see that if a PRS sequence is indefinitely repeated, the spectrum corresponding to the repeated sequence is "discretely white". The reverse is also true: if an indefinitely repeated sequence of pulses has a "discretely white" spectrum it is a PRS.

In order to see this we start from the property, quoted above, that the spectrum of the auto-correlation is the power density spectrum of the original function. In the present case of indefinitely repeated sequences $f(t)$ of length NT we would correspondingly have for their auto-correlation function $\phi(\tau)$ in terms of their Fourier components $F(\frac{2\pi q}{NT})$:

$$\phi(\tau) = \sum_{q=-\infty}^{q=+\infty} F\left(\frac{2\pi q}{NT}\right) F^*\left(\frac{2\pi q}{NT}\right) e^{\frac{j2\pi q}{NT}\tau} \quad (7.1.E1).$$

Obviously if the discrete spectrum $f(t)$ is white $F(\frac{2\pi q}{NT}) F^*(\frac{2\pi q}{NT})$ is in-

dependent of q and the sum of the exponentials can only add up to a sum different from zero if τ is a multiple of NT .

In summary we now have the following result:

The spectrum of an indefinitely repeated sequence of pulses of length NT is discretely flat if and only if the repeated sequence has an auto-correlation function $\phi(\tau)$ that is periodic with period NT and only different from zero if $\tau = nNT$.

The question remains whether PRSs as we have described them really exist. This has been shown to be the case. They can be generated by computers. After proper DC adjustment [6] they have the kind of auto-correlation function that we described in the above and constitute a limited subset of two valued sequences of length $NT = (2^M - 1)T$, where M is a positive integer [7]. Therefore N must be 1, 3, 7, 15, 31, 63, 127, 255, 511, etc.

7.2 A Single Unrepeated PRS as a GCR

After establishing that an indefinitely repeated PRS has a white discrete spectrum we now have to shift our attention to the case of an unrepeated PRS. We will do this by considering what happens if we add to the PRS an increasing number of pulses with zero magnitude, a process known as "padding".

Consider the indefinitely repeated PRS sequence

$$a_1, a_2, \dots, a_{(m-1)}, a_m \quad \text{cyclically repeated} \quad (7.2.E1).$$

If we start constructing its auto-correlation function by effecting a single shift of magnitude T we obtain

$$a_2, a_3, \dots, a_m, a_1 \quad \text{cyclically repeated} \quad (7.2.E2)$$

and the corresponding first member of the auto-correlation function is

$$a_1 a_2 + a_2 a_3 + \dots + a_{(m-1)} a_m + a_m a_1 \quad \text{cyclically repeated} \quad (7.2.E3).$$

Since our sequence is a PRS this must be zero since it is a member of the auto-correlation function.

If we "pad" one single zero onto the unrepeated sequence and then repeat the result we get

$$0, a_1, a_2, \dots, a_{(m-1)}, a_m \quad \text{cyclically repeated} \quad (7.2.E4)$$

and after a shift of one

$$a_1, a_2, a_3, \dots, a_{(m-1)}, a_m, 0 \quad \text{cyclically repeated} \quad (7.2.E5).$$

Now the corresponding member of the auto-correlation function is

$$a_1 a_2, + a_2 a_3, + \dots, + a_{(m-1)} a_m \quad \text{cyclically repeated} \quad (7.2.E6).$$

Compared to 7.2.E3 the term $a_m a_1$ has vanished and since a_m and a_1 both differ from 0 the result cannot be 0. For that reason the single-padded PRS is not a PRS anymore and as a consequence the spectrum though still discrete is not white.

Additional padding with n additional zeros will modify 7.2.E4 into

$$\begin{array}{l} \leftarrow n \rightarrow \\ 0 \dots 0, 0, a_1, a_2, \dots, a_{(m-1)}, a_m \quad \text{cyclically repeated,} \end{array}$$

so that 7.2.E5 becomes

$$\begin{array}{l} \leftarrow n \rightarrow \\ 0 \dots 0, a_1, a_2, a_3, \dots, a_{(m-1)}, a_m, 0 \quad \text{cyclically repeated,} \end{array}$$

again leading to 7.2.E6 for the correlation's member with shift T , and we have thus shown that all padding destroys the flatness of the spectrum. If we go to the

limit with infinite padding we have changed the indefinitely repeated PRS to a single PRS. We thus see that a single PRS has a non flat spectrum and therefore does not seem to be a good candidate for the role of a GCR. There is more to be said here:

The spectrum of an indefinitely repeated signal is, as is well known, a sampling of the spectrum of the unrepeated function, with components at frequencies that are multiples of the inverse of the length in time of the unrepeated sequence. Consequently padding only changes the sampling of the spectrum. In particular if we pad with a number of zeros that is a multiple of N the original spectral components are still present and they still constitute a white discrete subset of the spectrum.

We thus see that the spectrum of an unrepeated PRS, though not white, still contains the original "white" discrete components of the spectrum of the indefinitely repeated PRS. This means that we have to consider the issue of the spectral density of these "white" components in comparison with the spectral density that we would require for good echo cancellation. We will eventually see that the latter is larger than the former.

7.3 Permissible Length of an Unrepeated PRS and Spectral Density of Its "White" Spectral Components

Suppose we have a single unrepeated PRS of N pulses, where N is constrained to value $2^M - 1$ as discussed in 7.1. If we send this sequence in a bandwidth B , we know from the Nyquist theorem that the separation T of the

pulses has to satisfy the criterion that $T > \frac{1}{2B}$. Now we also have to fit the single sequence of duration NT onto a TV line duration minus the time required for sync and color burst. This would amount in NTSC to some 52 μsec . Therefore NT must be less than 52 μsec . If we now put $B = 4.2$ Mhz we find that N must be less than 437. The highest available value of form $2^M - 1$ thus is 255. The corresponding sequence duration would be $\frac{255}{437} 52 \mu\text{sec}$ or some 30 μsec . As we shall see the main significance of this for our purposes is that this is considerably less than the time range in which echoes tend to occur. We assumed in our actual system that this range extends from $-5 \mu\text{sec}$ to $+45 \mu\text{sec}$ but larger ranges have occasionally been observed. Let us accordingly assume for the present discussion that the range extends over at least 50 μsec . With a bandwidth of 4.2 Mhz and therefore a sampling frequency of at least 8.4 Mhz this range would require at least $420 + 255$ samples (where the 255 is only for the samples of the PRS itself). This minimum number of samples in the time domain can be translated in a minimum required spectral density $\frac{420 + 255}{2B}$ for proper cancellation. This is far more than the spectral density of $\frac{255}{2B}$ that can be attained with the "white" discrete components of the PRS. We therefore conclude that the "white" components will not be sufficient for echo cancellation and that the information contained in the rest of the spectrum is needed as well.

The situation may perhaps be clarified by considering what would actually happen when we indeed attempt cancellation, while trying to use exclusively the "white" components of the PRS spectrum.

To begin with we then have to realize that the echoed (single) PRS must of necessity have a duration that is larger than the duration of the original PRS. If

we want to restrict the signal processing to only the "white" discrete components of the PRS spectrum and at the same time take into account the full channel bandwidth the only way to give the echoed PRS spectral components at the same frequencies as the original (repeated) PRS is to take into account a portion of the echoed PRS with the same duration as the original PRS and to repeat that portion indefinitely † in order to have the DFT process in the receiver deliver a spectrum that is discrete and properly spaced. This means that some portion of the echoed PRS has to be ignored and that we at best hope to be able to cancel the remainder. We thus see that full restoration of the channel under these restrictions is not possible. In other words ignoring the non-white components of the spectrum of the PRS simply implies the impossibility of echo cancellation!

One could try to argue that a way out would be to take into account a period of time equal to a multiple of the duration of the original PRS. Indefinitely repeating this signal would yield frequencies at the location of the "white" components of the original PRS plus additional ones in between. Of course these would then have to be ignored, but again that would mean that we are canceling something else than the echo signals (also see the foot note below †).

Obviously this state of affairs worsens if the echo range becomes longer. Another way of stating this is that the situation is expected to improve to some degree when this range becomes shorter. That is the reason why GCRs with PRS character are being used in telephony. Here propagation occurs in cables or micro wave links, where echo delays are relatively short as compared with terrestrial TV. It will be clear from the above that still cancellation will not be complete unless

† Note that repetition would render distinction between pre- and post echoes impossible.

the echoed PRS has the same length as the original one, but that would mean that there are no echoes to cancel.

7.4 Illustrations

In order to demonstrate what was said in the above, we present pictures as obtained by computer calculations of various spectra. They were obtained by us-

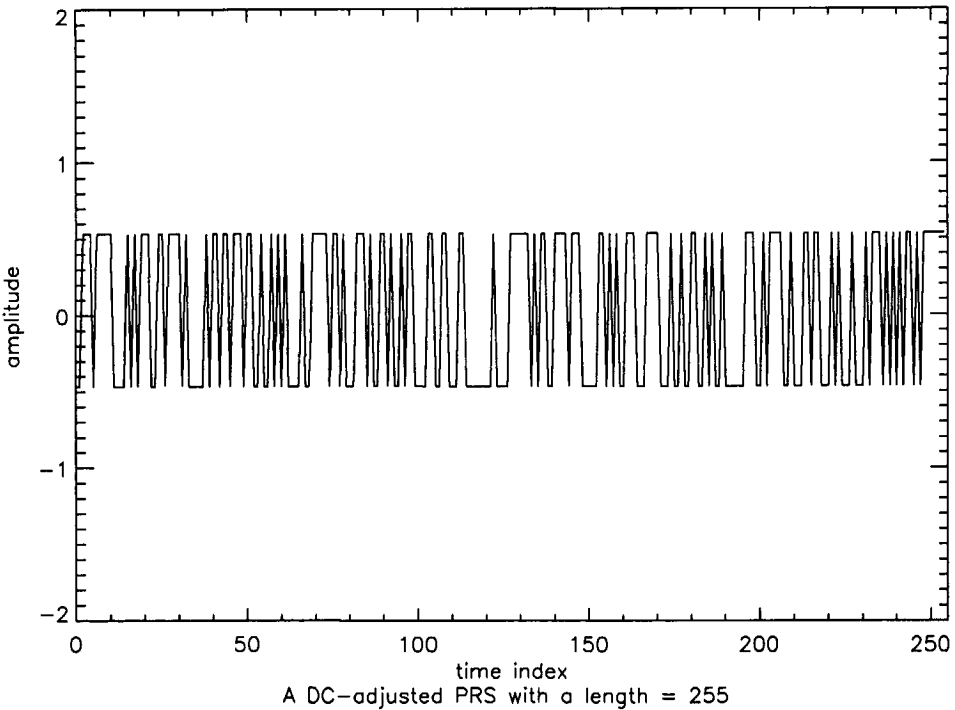


Figure 7.4.F1

ing various DFT and FFT packages.

Figure 7.4.F1 shows one of the possible PRSs with length $2^M - 1$ where $M = 8$. In this figure we see single pulses (sharp tips) as well as small sequences of several identical pulses (flat tips). The PRS has been DC-adjusted as described in 7.1. Of course the PRS is shown without repetitions.

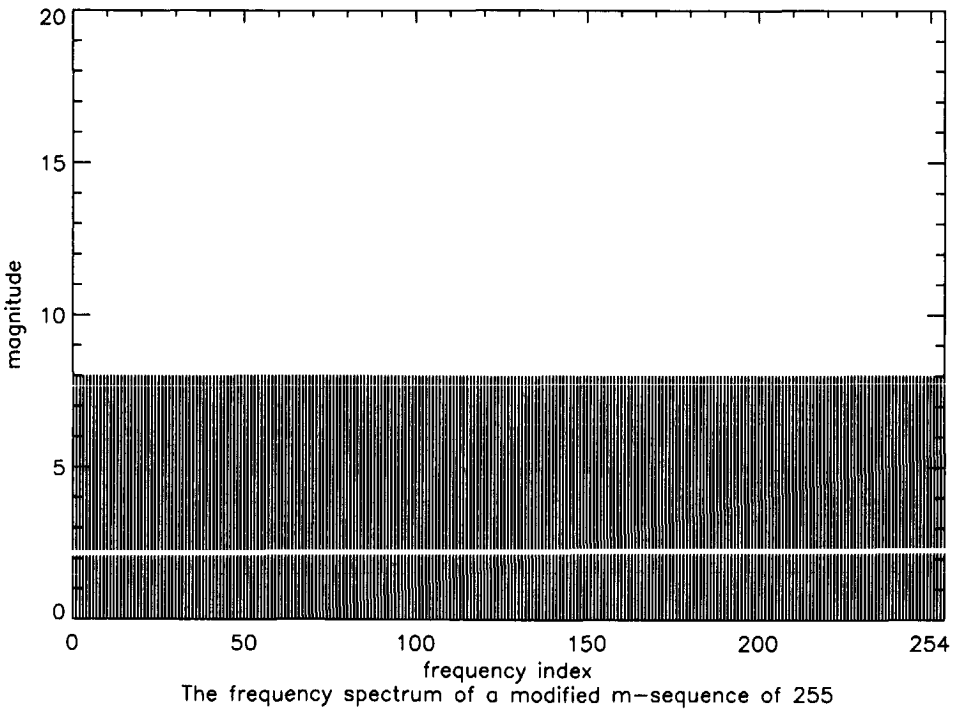
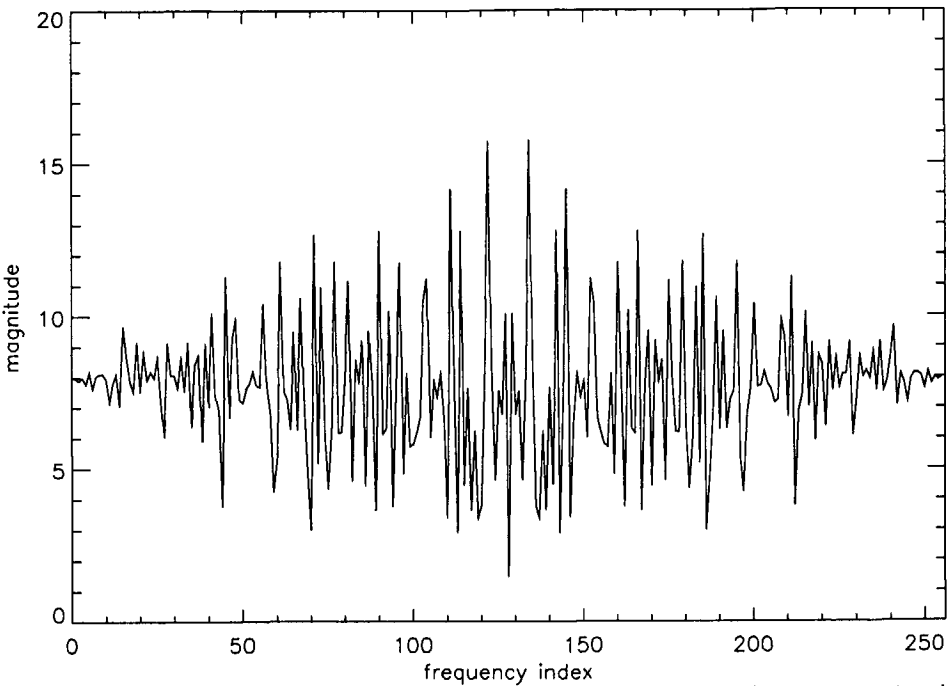


Figure 7.4.F2

Figure 7.4.F2 shows the discrete spectrum of the same sequence. Clearly the

spectral density is constant, as predicted by the considerations in the present chapter. Note that along the vertical axis the magnitude of the spectral components is plotted. Their phase is not shown, since it is immaterial in this context.

Figure 7.4.F3 shows the spectrum of the same PRS, padded with only one extra zero. As can be clearly seen the spectral density varies wildly. This is still more the case in the spectrum shown in Figure 7.4.F4 where instead of only one



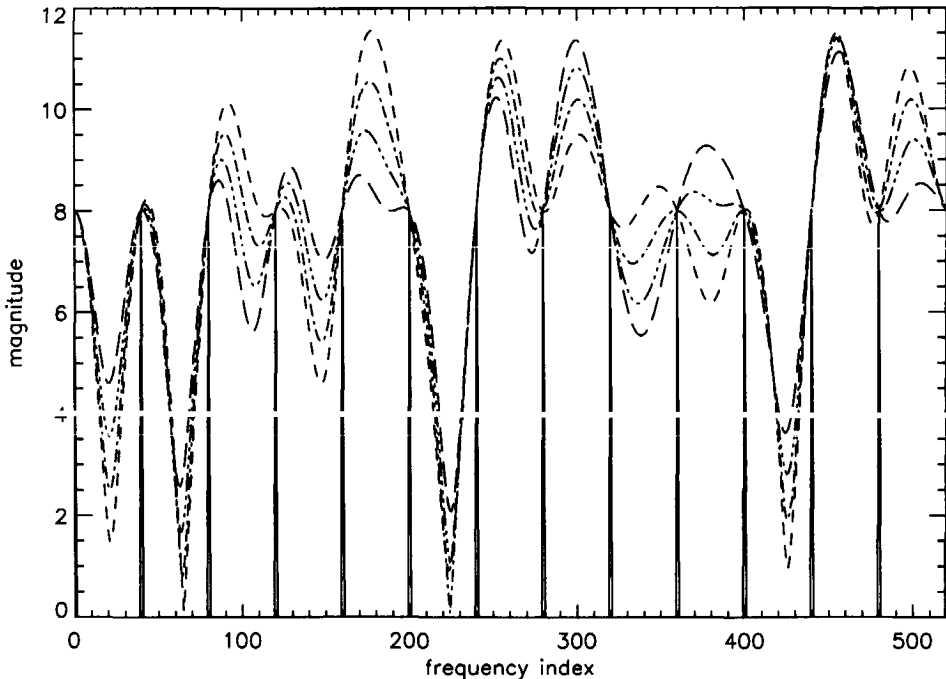
The frequency spectrum of a DC-adjusted PRS of length 255 with 1 zero extension

Figure 7.4.F3

zero a series of 9945 zeros has been added. Since 9945 is a multiple of 255 the flatness of the original (unpadded) discrete spectrum has remained, as can be seen

at frequency indexes 40, 80 etc. It should be noted that this figure only shows the first 520 spectral components of the total of 40 times 255, or 10200. Note also that in this case very deep notches are present. The difference between the various curves is the location of the padding, which has been added after rotating the original sequence by 1, 2, 3 and 4 samples respectively. As expected this does not affect the original flatness of the discrete spectrum components.

All spectra shown in this chapter are periodic because we are looking at a series of infinitely narrow pulses, i.e. at a sampled function. The pictures show the magnitude of the spectral components between zero frequency and twice the Nyquist frequency.



The spectra of rotated (1,2,3,4 samples) PRS of length 255 with 9945 zero extensions (zoomed)

Figure 7.4.F4

7.5 Conclusions

In the above we have shown that the spectrum of an unrepeated PRS is not flat to the extent required by proper echo cancellation in television. Obviously the use of a repeated PRS would only increase the non flatness of the spectrum. Therefore PRS signals have to be discarded as GCRs.

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† RFP: Request For Proposal.

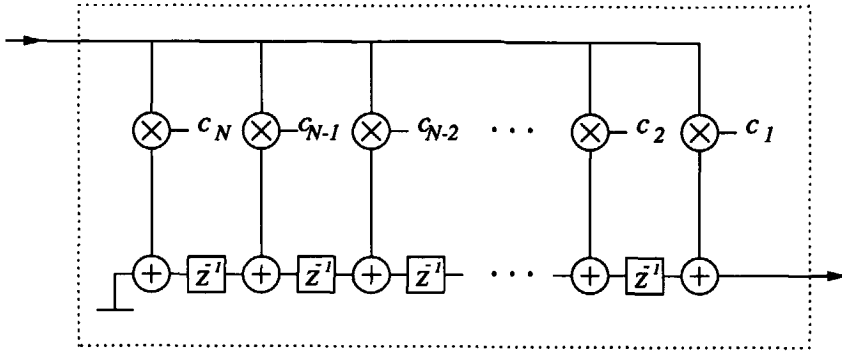
CHAPTER 8

ALGORITHMS FOR CHANNEL CHARACTERIZATION

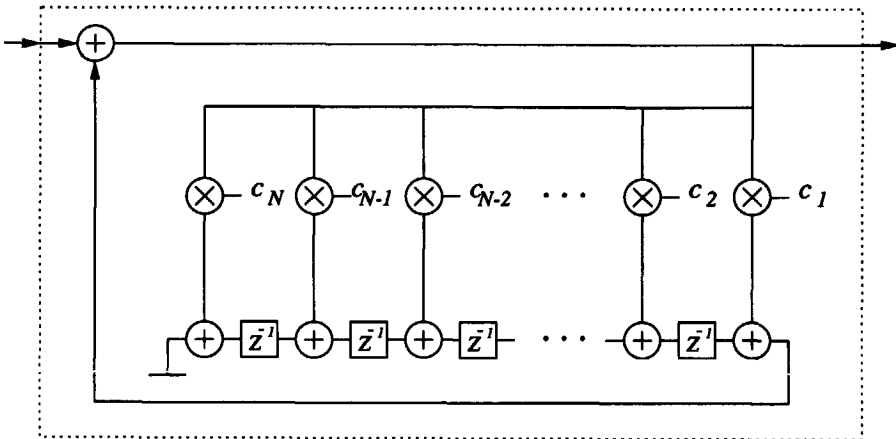
In the previous chapters we presented the reasoning for the design of optimal GCRs for the various TV systems that are in use in the world. The next steps would naturally be (1) using these GCRs to characterize the channel and (2) using the channel characteristics thus obtained to perform echo cancellation. In practice this procedure of characterization and subsequent cancellation is not optimal and one has to iterate between the two processes. The main reason for this will be explained in the following.

8.1 Interaction between Channel Characterization Echo Cancellation

Clearly the implementation of echo cancellation as a whole can only be done economically by digital methods, in particular by using programmable digital filters with filter coefficients that are calculated and adjusted continuously on the basis of calculations by a microprocessor in the TV set. Now these digital filters basically can be either of the linear "finite impulse response" (FIR) type or of the feed-back "infinite impulse response" (IIR) type, both illustrated in Fig. 8.1.F1.



(a) *The Structure of A FIR Filter*



(b) *The Structure of An IIR Filter*

Fig. 8.1.F1

The Filter Structures of a FIR Filter and an IIR Filter

As will be clear from the figure an IIR filter's response emerges at its output as soon as there is an input, due to the direct connection between input and output, no matter how the FIR that it contains has been programmed. For this reason an IIR cannot be used for the cancellation of pre-echoes. These in principle can be

canceled however with the aid of a FIR filter.

On the other hand a FIR filter if used for echo cancellation in principle produces a delayed and attenuated version of the echoes it cancels (see 1.3.3.2), and in order to cancel these in turn one has to increase the length of the filter. The process then repeats itself and consequently for (infinitely) complete cancellation one needs an (infinitely) long FIR.

Another difference between FIR and IIR filters is that the feedback character of an IIR implies that apart from a simple correction it acts as the inverse of the FIR in its feedback loop. Consequently all that is needed for cancellation is to have the impulse response of the channel further processed (see below) and copied into its coefficients. This is in sharp contrast with the fact that elaborate adaptive algorithms have to be carried out by the microprocessor to program a FIR filter for cancellation.

Summing up the situation now, it is like this:

Although it seems straightforward and therefore simple to just characterize the total channel and to program a single filter on that basis to remove the distortion that the channel causes, this is in fact not practical. If the filter is an IIR pre-echoes cannot be canceled, if it is a FIR it would have to be very long and thus costly in terms of hardware and the required "on the fly" processing time for its programming, see 9.1, by adaptation.

So we have to use both kinds of filters, in principle a FIR for the pre-echoes and an IIR for the post-echoes. Here we luckily can benefit from the fact that pre-echoes tend to occur within only 5 μ sec before the main signal while post-

echoes can arrive as late as 45 μ sec after the main signal.

This gives us the opportunity to use a short FIR to cancel the pre-echoes on the basis of adaptation as described in the next chapter. Then we can set up a new channel model consisting of a concatenation of a channel model that we obtain as described below and the FIR filter. This concatenated model will not show pre-echoes anymore and thus can be used for the design of the subsequent IIR filter which will complete the cancellation.

8.2 Characterization by Frequency Division

We can be very short on this topic: it was already treated in 5.1 in connection with finding the requirements that a GCR has to satisfy. From 5.1.E2 we see that a GCR that contains enough energy for all frequencies allows us to assume

$$H(\omega) = \frac{R(\omega)}{G(\omega)} \quad (8.2.E1),$$

where $H(\omega)$ is the frequency characteristic of the channel, $R(\omega)$ is the Fourier transform of the received distorted GCR and $G(\omega)$ is the Fourier transform of the GCR in pure form.

Multiplication of the received video signal by $\frac{1}{H(\omega)}$ would then cancel the echoes as well as other linear distortions automatically.

In our implementations we did not use this method, because of its sensitivity to noise as described in 5.1. Workers in NHK and NEC [1] have studied and used it with some success however.

One of the chief merits of the method is that it is simple in terms of required processing power and provides some of the insights that we presented in previous chapters.

8.3 Characterization by Correlation

Also this method was treated in detail previously (5.2) in order to demonstrate the advantages of a GCR with a flat spectrum. It was shown there that if the GCR is a real signal with a flat spectral energy distribution, the cross-correlation of the received distorted GCR and the GCR in pure form is identical to the impulse response function of the channel. The validity of that derivation is not affected if we extend the bandwidth of the GCR somewhat beyond the bandwidth of the channel. This would move the roll-off out of the channel, thus increasing the flatness of the spectrum within the channel.

This is the method that we used in most of our work. The flat spectrum of our family of GCRs gave us the unique opportunity to do so. Moreover it constitutes an attractive trade-off between "simple, but noise sensitive" (frequency division) and "robust to noise, but complex" (adaptation as described below).

8.4 Characterization by Adapting the Coefficients of a FIR Filter

Let us begin this section by explicitly stating that its content is well known. In fact what we present here is a short description of an adaptation procedure commonly referred to as the Least Mean Square (LMS) adaptation method [2]. We did not use it for characterization in our work as explained in 8.3, but we did use a closely related procedure for cancellation as will be discussed in the next chapter.

Assume that we want to characterize a channel $h(t)$ by creating a FIR filter with N coefficients $c(n)$, $n = 1, \dots, N$. We would do this by sampling the received distorted GCR resulting in M samples $r(n)$, $n = 1, \dots, M$, which puts a lower limit on M in order to avoid aliasing.

Our task now is to determine the $c(n)$, in such a way that if an undistorted GCR $g(n)$, $n = 1, \dots, N$ is input into the filter with coefficients $c(n)$ the output $s(n)$ would equal $r(n)$, which would mean that the filter has become identical to the channel. We will do this by successive approximation.

Obviously

$$r(n) = \sum_{l=1}^n h_l g_{n-l+1}, \quad l \leq n \quad (8.4.E1)$$

and

$$s(n) = \sum_{l=1}^N c_l g_{n-l+1}, \quad l \leq n \quad (8.4.E2).$$

Thus the "error" $e(n)$ that occurs in the various samples at the output of the filter would be

$$e(n) = \sum_{l=1}^N (h_l - c_l) g_{n-l+1}, \quad (8.4.E3).$$

A sensible way [2] to minimize the total joint errors over all n , though not at all the only possible one [3], is to sum the squares of these errors over all samples of the output and then to minimize the sum.

The number of samples to be taken into account in this summation is larger than N . In fact it is easy to establish that if the GCR would on account of its length have M samples $g(n)$ at the same sampling rate as used for $r(n)$ and $s(n)$, the number of samples at the channel output would be $M + N - 1$. So this would also be the number of samples to be taken of $s(n)$.

This leads immediately to a total of squared errors amounting to

$$E = \sum_{n=1}^{M+N-1} \left[\sum_{l=1}^N (h_l - c_l) g_{n-l+1} \right]^2 \quad (8.4.E4)$$

It is not difficult to minimize this. We first calculate the derivative of E with respect to each $c(n)$

$$\begin{aligned} \frac{\partial E}{\partial c_l} &= -2 \sum_{n=1}^{M+N-1} \left[\sum_{l=1}^N (h_l - c_l) g_{n-l+1} \right] g_{n-l+1} \\ &= -2 \sum_{n=1}^{M+N-1} e_n g_{n-l+1} \end{aligned} \quad (8.4.E5).$$

Now imagine that in $c(n)$ space we want to make a step of fixed length $\vec{\Delta c}$ in a direction that maximizes the decrease in E it would produce. We would have

$$\begin{aligned} \Delta E &= \sum_{l=1}^N \left(\frac{\partial E}{\partial c_l} \right) \vec{\Delta c}_l \\ &= \text{grad}_{c_l} E \cdot \vec{\Delta c} \\ &= \left| \text{grad}_{c_l} E \right| \left| \vec{\Delta c} \right| \cos \phi \end{aligned} \quad (8.4.E6)$$

where ϕ is the angle in multi dimensional $c(n)$ space between $\text{grad} E$, with components 8.4.E5 and the direction of the step $\vec{\Delta c}$. Consequently the increase in E when we move in that direction is maximized if ϕ is 0, i.e. when we move in the direction of the gradient. The decrease in E then is maximized when we move in the opposite direction: this is the so-called direction of steepest descent.

If we take steps like this repeatedly we will have successive versions \vec{c}^k , $k = 1, 2, \dots$ of the vectors with components $c(n)$, $n = 1, \dots, N$ as follows

$$\vec{c}_l^k = \vec{c}_l^{k-1} + 2\mu \vec{e}^k \times^l \vec{g}^{k-1} \quad (8.4.E7)$$

where

$$\vec{e}^k \times^l \vec{g}^{k-1} = \sum_{n=1}^{M+N+1} e_n^k g_{n-l+1}^{k-1} \quad (8.4.E8)$$

We added in the above equations a superscript to \vec{g} in order to preserve consistency with the literature [2] in spite of the fact that in the present context \vec{g} is the same for all iterations.

Here the number μ has been introduced in order to have an adjustable parameter during the minimization process of E . If μ is too large the steps are too large to get close to the minimum. Instead the process will "jump" over the minimum, then jump back and thus become unstable. If μ is too small the process will become too slow. Clearly a trade off has to be made here. One can even have bigger initial steps and smaller final steps, or detect back jumps and then decrease μ .

We note in conclusion that the physical meaning of 8.4.E8 is that in each successive approximation the correction to $c(l)$, $l = 1, \dots, N$, is 2μ times the correlation between the signals $e(n)$ and $g(n)$ at a relative delay of $l - 1$ samples in $g(n)$. Since in our calculations in the above l runs from 1 to N this delay is equal to or larger than 0: negative delays are excluded. At any rate we see that for each iteration the correction to the tap values $c(n)$ can be made since all the information needed is available to the microcomputer. It seems to make sense to make the initial setting of the tap values independent of the local echo constellation in the channel that is used, since it would have to be programmed into the TV

set when it is manufactured. The most obvious choice would be to assume the channel to have no echoes at all, that is to assume that the impulse response function of the channel is a band limited delta pulse which has the shape of $\frac{\sin(x)}{(x)}$. Correspondingly the initial setting of the $c(n)$ would be a suitably windowed version of that function so that it would occupy N sample moments.

This shows that there is a trade off to be made here between complexity (large N) and cost (lower N). This also shows that it cannot be determined from the above calculation as such what the attainable limit for E is going to be, even if it is assumed that channel noise and quantization noise are zero.

8.5 Considerations on Choosing a Channel Characterization Method

We refrained from using the frequency division method since in all cases $H(\omega)$ obviously goes to zero at the edges of the channel and sometimes can be very small in a "notch" in the spectrum. This makes the division inaccurate and creates problems of noise and even instability. Furthermore computation-wise the method is not clearly the best of the methods mentioned.

Adaptive characterization of the channel is cumbersome and therefore costly in computation effort and was discarded in our practical work for that reason although the method is robust against noise.

The method preferred by us was correlation. It is computationally simple and almost by definition insensitive to noise while it uses the rather unique property of our GCR that it has a flat spectrum.

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CHAPTER 9

PROGRAMMING OF FILTERS FOR ECHO CANCELLATION

In this chapter we will finally embark on the subject of actual echo cancellation with linear digital programmable filters. As already stated in the previous chapter we will have to consider a combination of a FIR filter and an IIR filter. The latter contains in its feedback loop another FIR filter which we shall refer to as FIR1 to avoid confusion. The FIR filter will be used to cancel pre-echoes. This is something which an IIR filter cannot do because no matter how it is programmed it will produce an output as soon as it receives an input. The IIR filter will have to cancel the post-echoes. We will show below (in section 9.2) that for the proper determination of its coefficients we need the channel characteristic as well as the values of the coefficients that make the FIR filter remove the pre-echoes. The latter can be determined (see section 9.1) by the use of an adaptation method [1] that is very similar to the one described in the previous chapter in connection with channel characterization.

9.1 Adaptive Algorithm for Finding the FIR Coefficients for Pre-echo and Nearby Post-echo Cancellation

The FIR filter that we have to construct by adaptation here, has to convert a

distorted input into an ideal, that is undistorted, output. We can in principle choose any input signal that the actual channel delivers to be converted to the corresponding output of an ideal, band limited, channel. Obviously two reasonable choices for the input are the channel's impulse response together with the impulse response of an ideal band limited channel or the distorted GCR as received by our receiver together with the stored original GCR. The first choice requires knowledge of the impulse response $h(t)$ of the channel as obtained from the channel characterization procedures described in chapter 7. We would expect $h(t)$ to be shorter than the distorted GCR and therefore it seems to be the more economical choice in terms of computational resources required. The corresponding output to be realized by the FIR would then be a band limited $\delta(t)$ pulse, suitably time windowed again. This is the course that we are going to adopt in the following. It should be remembered here that in doing so we fully use the benefits of our GCR design as described in chapters 5 and 6 since our channel model $h(t)$ has been obtained from a comparison in one of the forms treated in chapter 8 of original and distorted GCR.

If the time windowed band limited $\delta(t)$ pulse that serves as our ideal output is centered at $t = 0$ we will have to consider the input only from a suitable negative time, to be determined from statistics on actual pre-echoes, up to a suitable positive time which would take care of cancellation by the FIR of nearby post-echoes. Let this require M samples and let the FIR to be programmed have N taps. If the tap values are $c(n)$, the output of the FIR will be given by

$$y(n) = \sum_{l=1}^N h(n-l+1) c(l) \quad (9.1.E1)$$

So, if the stored samples of our time windowed $\delta(t)$ pulse are $u(n)$, the error $e(n)$ in $y(n)$ is

$$e(n) = u(n) - y(n) \quad (9.1.E2)$$

Therefore the sum of the squares of all $M + N - 1$ samples of the error is

$$E = \sum_{n=1}^{M+N-1} [u(n) - \sum_{l=1}^N h(n-l+1) c(l)]^2 \quad (9.1.E3).$$

This equation enables us to form, as a digression, a mental picture in three dimensions of the multidimensional adaptation process to be described below in an algebraic way. We will assume that we are programming a filter with only two coefficients (c_1 and c_2), so the three dimensions to be considered are E , c_1 and c_2 and E can be written, from 9.1.E3 as

$$E = P + Oc_1 + Rc_2 + Sc_1^2 + Tc_2^2 + Uc_1c_2$$

where S and T are sums of squares and thus positive. Consequently E must have a single minimum as a function of c_1 and c_2 and therefore should look like the quadratic surface depicted in Fig 9.1.F1. At the start of the adaptation process we are somewhere on this surface and we have to make a steepest descent towards the minimum. We do this by determining the direction in which we have to travel in

the c_1, c_2 plane and then take a step of predetermined length in that direction staying on the E surface. If the step is so small that we did not jump over the minimum we have made progress and can repeat the procedure. If the step is too big we jumped over the minimum and might do the same again if we repeat the process without reducing the length of the step. This has to be done in the software of the system and we will return to this issue in Chapter 12, where this software is the subject.

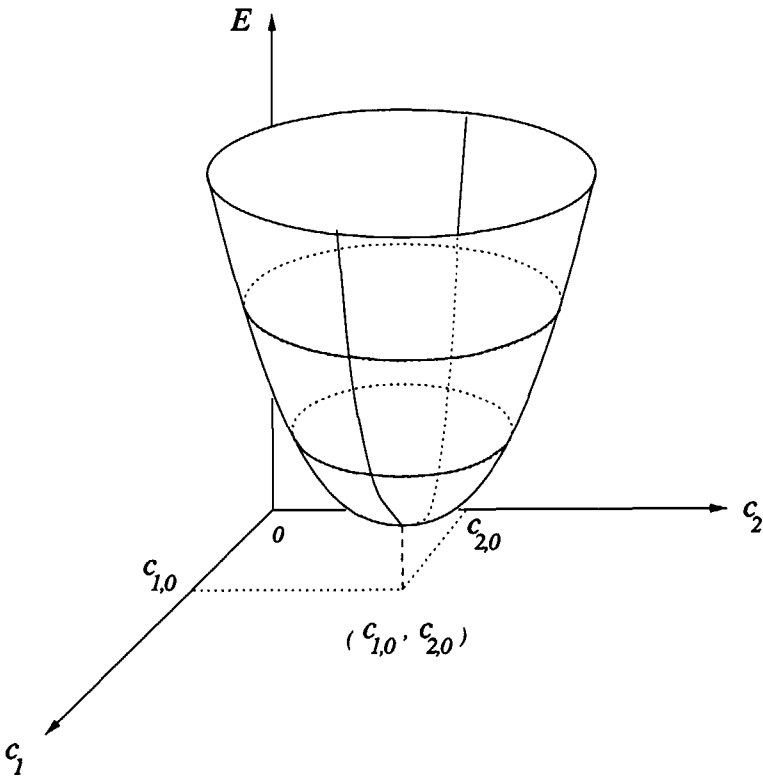


Figure 9.1.F1
Error E versus Coefficients C_1 and C_2

Returning now from the digression to our multidimensional treatment, we note from 9.1.E3 that the derivative of E with regard to $c(l)$ is

$$\frac{\partial E}{\partial c(l)} = \sum_{n=1}^{M+N-1} -2 e(n) h(n-l+1) \quad (9.1.E4).$$

As in the adaptation process described in Chapter 8 this indicates the direction of steepest ascent, from which the direction of steepest descent can be obtained by removing the minus sign. Introducing the adjustable factor μ again this yields

$$c^{k+1}(l) = c^k(l) + 2 \mu \sum_{n=1}^{M+N-1} e(n) h(n-l+1) \quad (9.1.E5),$$

where the superscripts $k+1$ and k indicate consecutive generations of coefficients.

This shows that correlation of the error vector $e(n)$ with the channel response function in each generation of filter coefficients yields the correction that has to be applied in order to arrive at the next generation.

In our actual echo cancelers we have used this method [1] for programming the FIR filter. As described in the following section the *FIR* 1 filter in the loop of the IIR filter could then be obtained by using the convolution of the FIR with the channel characteristic.

As in all adaptation processes we will have to initialize the $c(n)$ values in some way. A reasonable choice is to take a time windowed band limited $\delta(t)$ function.

9.2 The Relationship between the IIR Filter and the Combination of the Channel and the FIR Filter

Referring to Figure 9.2.F1 we consider the channel, represented by its frequency response $H(\omega)$ in series with the FIR filter $H_{FIR}(\omega)$ that serves for the removal of the pre-echoes and the IIR filter. The latter contains in its feed-back path another FIR filter with frequency response $H_{FIR1}(\omega)$ so that its frequency response is $\frac{1}{(1 + H_{FIR1}(\omega))}$. Consequently the frequency response of the total system shown in Figure 9.2.F1 is

$$\frac{H(\omega) H_{FIR}(\omega)}{(1 + H_{FIR1}(\omega))} \quad (9.2.E1).$$

Since the GCR has been designed to fit within the (limited) bandwidth of the channel and since the actual video signal also fits within this bandwidth we can state that in the case of perfect rendition of the original signals, without echoes and other linear distortions, the expression 9.2.E1 should be equal to 1. Consequently we must have [2]:

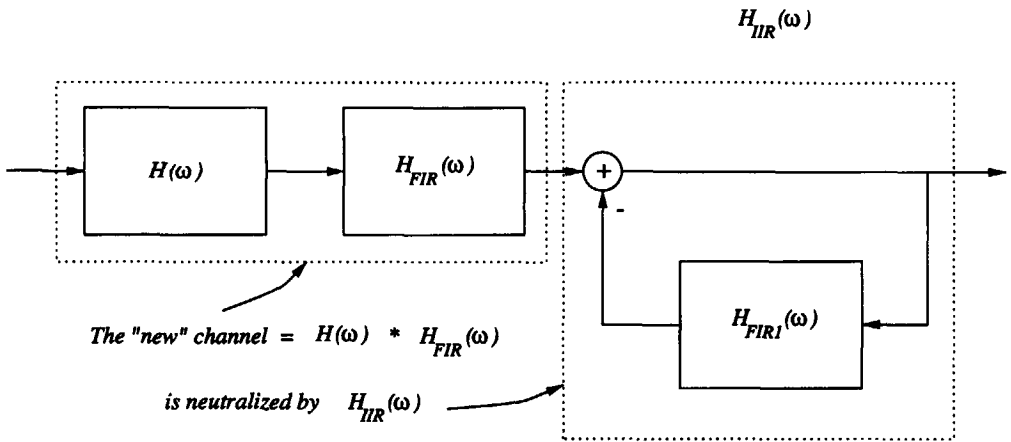


Figure 9.2.F1
The Block Diagram of the Echo Channel H
and the Correction Filters FIR and IIR

$$H_{FIR1}(\omega) = H(\omega) H_{FIR}(\omega) - 1 .$$

In order to arrive at a time domain description we have to Fourier transform this expression, which results in:

$$h_{FIR1}(t) = h(t) * h_{FIR}(t) - \delta(t) \quad (9.2.E2).$$

Here the * symbol as usual stands for a convolution. So $h_{FIR1}(t)$ can be obtained by first programming $h_{FIR}(t)$ to cancel the pre-echoes and (in practice) some near-by post-echoes and then convolving the filter thus constructed with the channel characteristic. If we then subtract a $\delta(t)$ function from the result we have found the $h_{FIR1}(t)$ that can be used to cancel the remaining echoes.

There remain a few practical problems. One is caused by the fact that in an actual echo canceler we will have to use a finite sampling rate and we will thus be unable to handle a delta function. This requires some approximation. The most sensible one seems to be to use a bandlimited delta function which is suitably time-windowed in accordance with hardware limitations that ensue from cost/performance considerations. These also apply in the realization of the convolution of $h(t)$ and $h_{FIR}(t)$: we have to limit the length in the time domain of both under the guidance of statistical evidence as to the occurrence of pre- and post-echoes.

Practical trade-offs of this nature as we (successfully) made them in our actual cancelers will be treated in chapters 11 and 12.

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CHAPTER 10

OVERALL CANCELLATION SYSTEM AND SIMULATION

In this chapter we will first give an overview of the whole process of channel characterization and echo cancellation as we described it up to now. Details about the hardware system that we actually built will be given in terms of both hardware and software in chapters 11 and 12. In the present chapter we shall be concerned only with the algorithmic side of the process, including our computer simulations of it.

Section 10.1 gives the overall system diagram. It does not present new information but merely sums up in a pictorial way what was said before.

Section 10.2 is a presentation of the main results of the simulations.

10.1 Main Block Diagram of the Software

Figure 10.1.F1 contains the overall block diagram. The various boxes have been numbered and will now be discussed in the order of their numbering. The input is, for the purpose of the present discussion, the GCR as received through the

channel. The output is the restored GCR. For the sake of clarity we will stick to the nomenclature and notation used in 9.2 throughout the discussion.

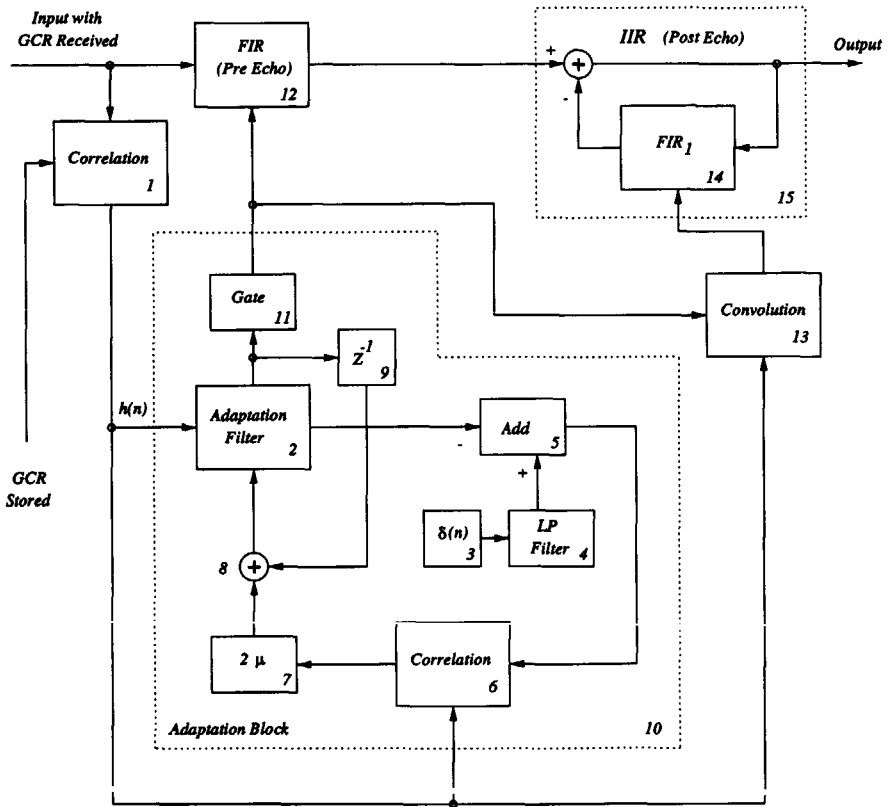


Figure 10.1.F1

Main Block Diagram of the Software for the Echo Cancellation System

The input is fed into the correlator (Block 1) together with the stored GCR, for channel characterization as discussed in chapter 8. The resulting channel

characteristic $h(n)$ is fed into the "software" adaptive FIR filter (Block 2) that serves throughout the process to calculate generations of sets of filter coefficients that would restore the pre-echo part of the distorted output of the channel (this is $h(n)$, assuming the input to the channel to be a delta pulse) to the ideal output (in this case a bandlimited delta pulse). This ideal output, is obtained from Block 3 (a delta pulse generator) and box 4 (a lowpass filter) for comparison with the output $y(n)$ of the filter in block 2 and on the side serves for the initialization of the filter coefficients of block 2 as discussed at the end of 9.2. This comparison (by subtraction of $y(n)$ and the bandlimited delta pulse) is done in Block 5. The result $e(n)$ is fed into Block 6 where correlation with $h(n)$ as described by 9.1.E5 takes place. The correlation function is multiplied in Block 7 with an adjustable constant μ to be added in Block 8 to the coefficient vector $c(l)$ of the filter as delivered by the previous cycle of the adaptive process. The link between the present and the previous cycle is represented in the figure by the delay of Block 9. The updated filter coefficients of (as default) the 16th generation (see 12.2) are gated out in box 11 and fed into another (definitive) FIR filter (Block 12), that has the received GCR as input and now removes the pre-echoes from it (see chapter 9). Convolution of the filter coefficients of this filter with $h(n)$ occurs in Block 13. The result constitutes the coefficients for the filter FIR1 that resides in the loop of the final IIR filter and that now can complete the cancellation process by removing the post-echoes.

10.2 Simulation Results

This section will conclude the theoretical part of this thesis by means of computer simulation results that will serve as a verification of the validity of our GCR

design as well as of the notions on channel characterization and echo cancellation that were presented in the previous texts.

We will do this by going through a series of figures that illustrate the state of affairs through the respective stages of the signal processing that we described.

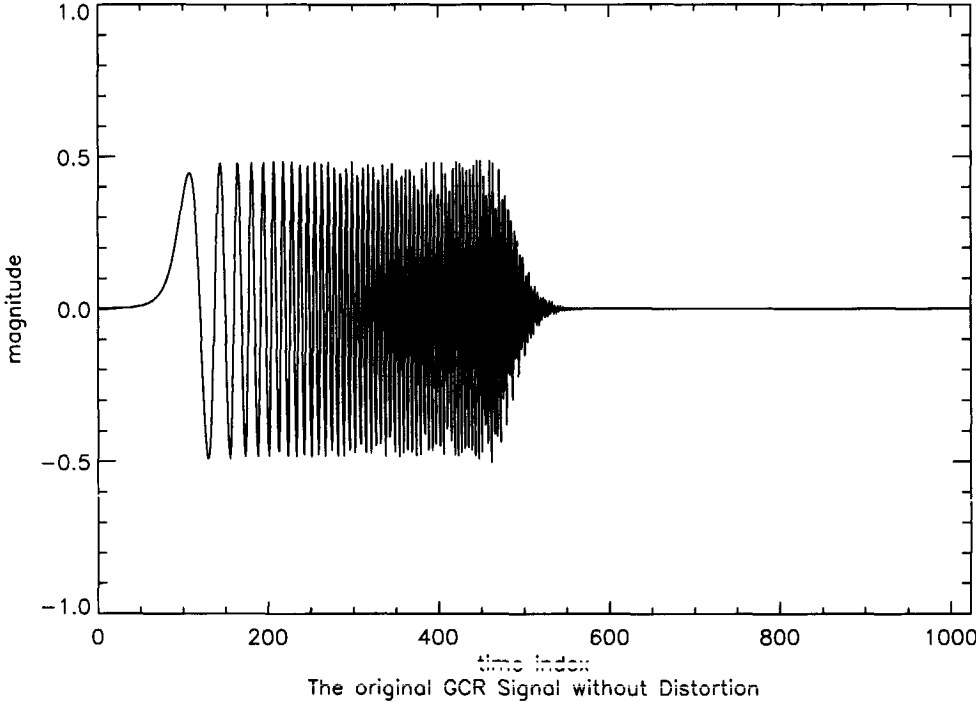


Figure 10.2.F1

Figure 10.2.F1 shows again a plot of the GCR in the time domain as it is sent and as it is stored in the memory of the receiver. Note the constant amplitude, the increase of frequency from beginning to end and the effect of the time windowing that we described.

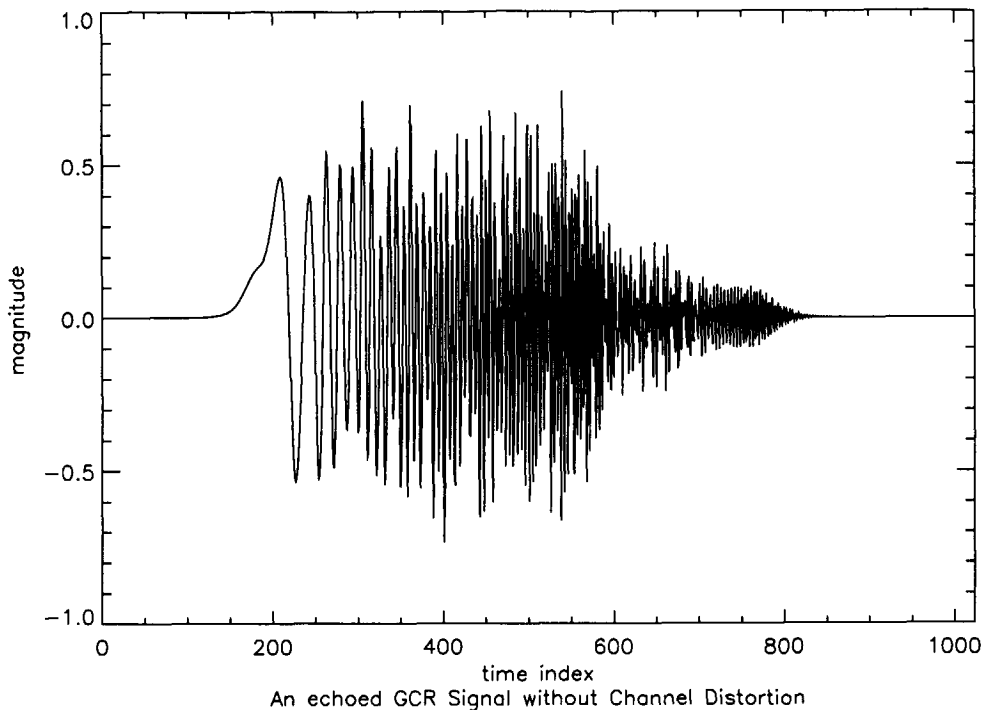


Figure 10.2.F2

Figure 10.2.F2 shows an echoed GCR as received. Note the fact that the amplitude of the echoed GCR is not constant. A particular choice was made of course as to the echo pattern in the (simulated) channel, which is shown as impulse response in Figure 10.2.F3. Pre-echoes as well as post echoes can be seen there, together with the main signal which is represented by the large peak at time index 0.

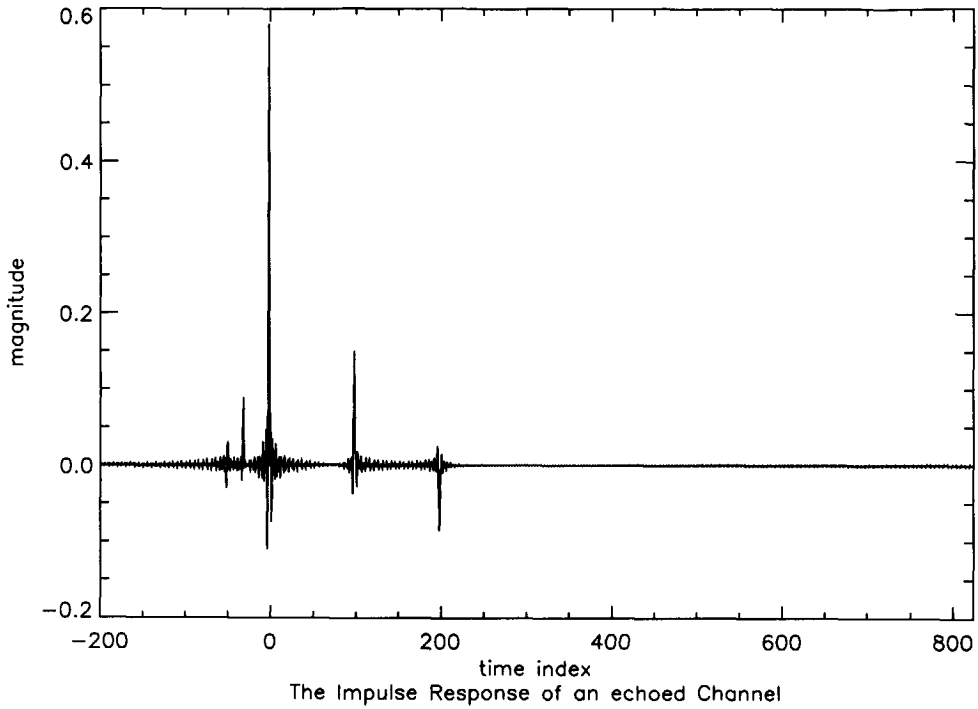


Figure 10.2.F3

Figure 10.2.F4 shows the coefficients of the FIR filter after application of the adaptation algorithm described in the previous section. Note the different time scale. We clearly see the influence of the pre-echoes as well as the tails of the (bandlimited) main signal at small negative and positive delays. The post-echoes at larger positive delays are not considered here as explained earlier in this chapter.

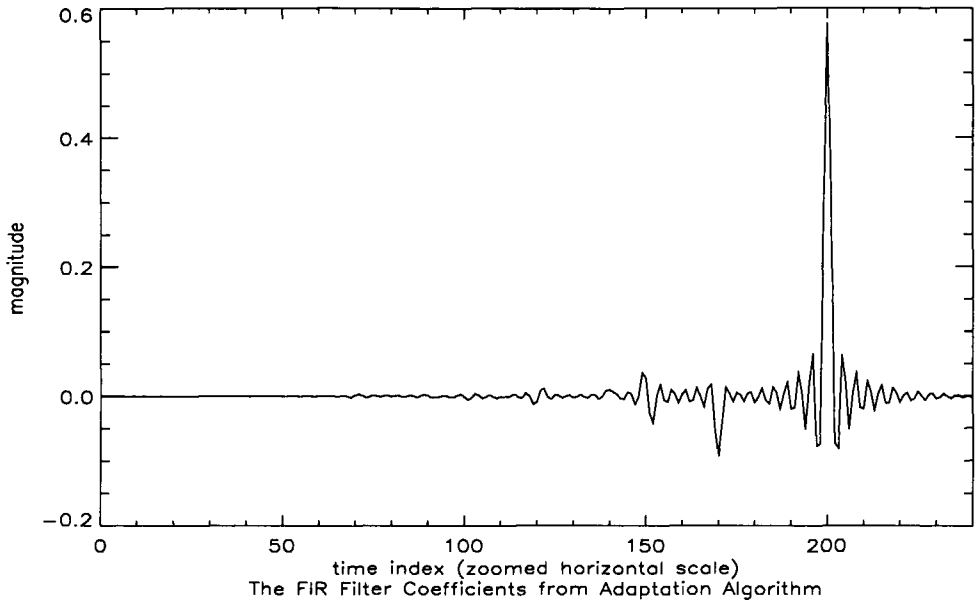


Figure 10.2.F4

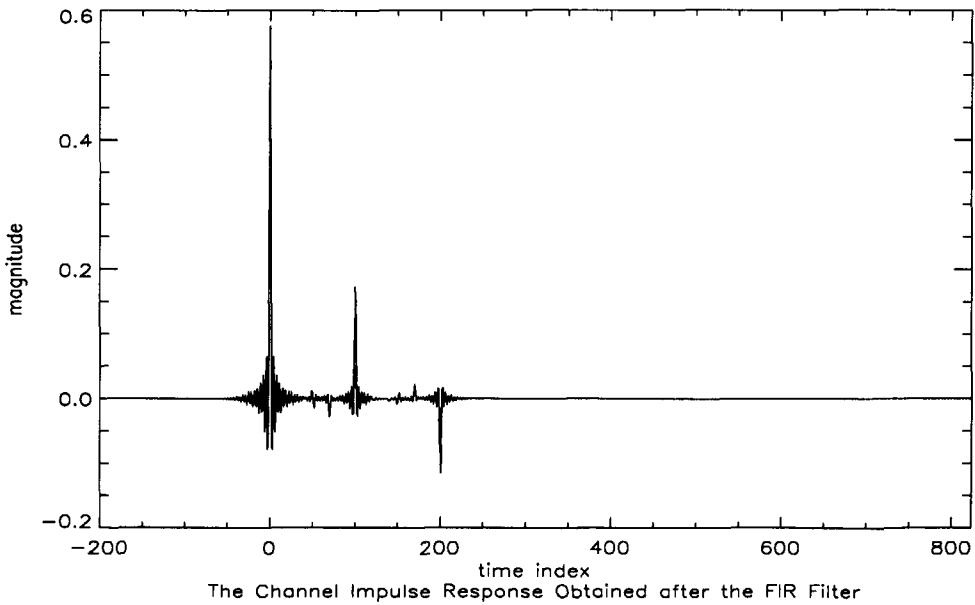


Figure 10.2.F5

Figure 10.2.F5 shows the impulse response of the channel, concatenated with the FIR filter: the latter now has eliminated the pre-echoes. The main signal as well as the post-echoes are left. The IIR filter coefficients should be a replica of Fig 10.2.F5 after subtraction of a bandlimited delta function as described in 9.2. This is shown in Figure 10.2.F6.

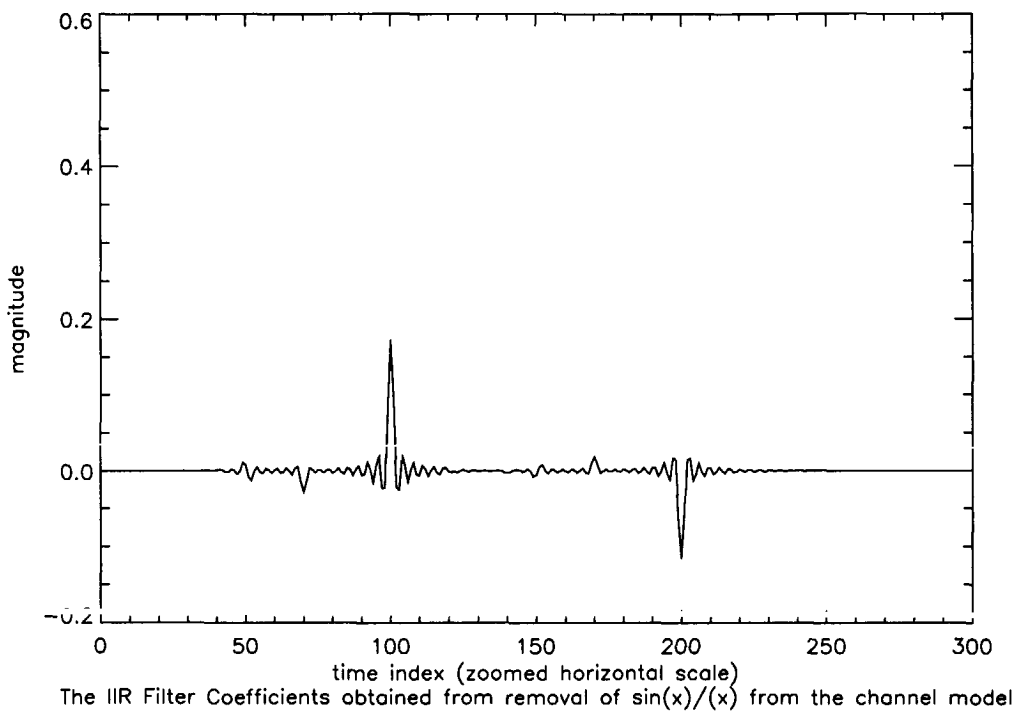


Figure 10.2.F6

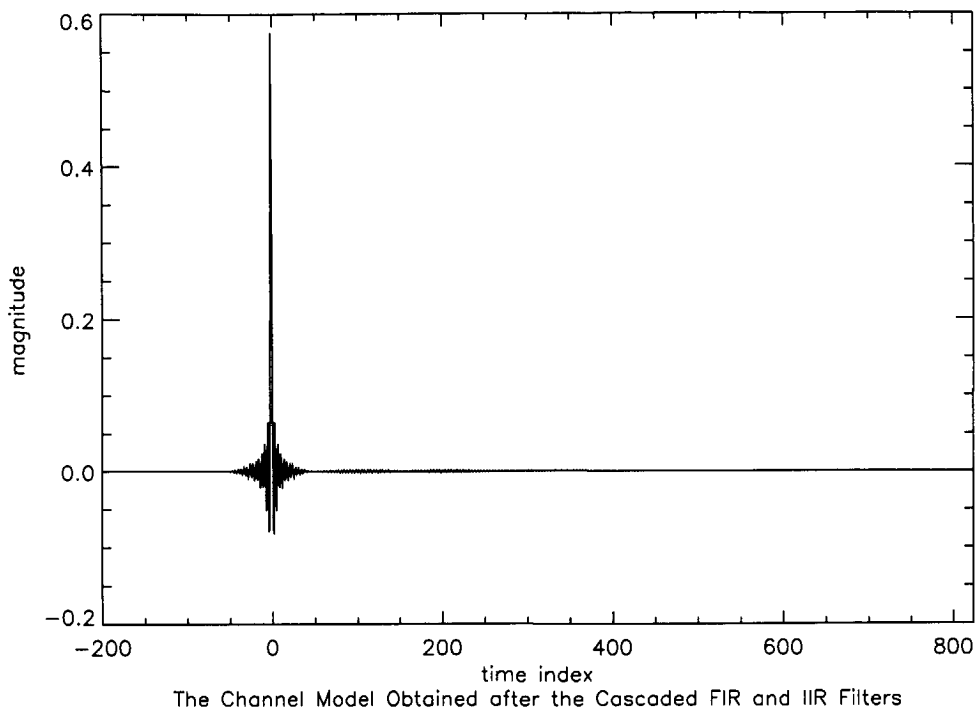
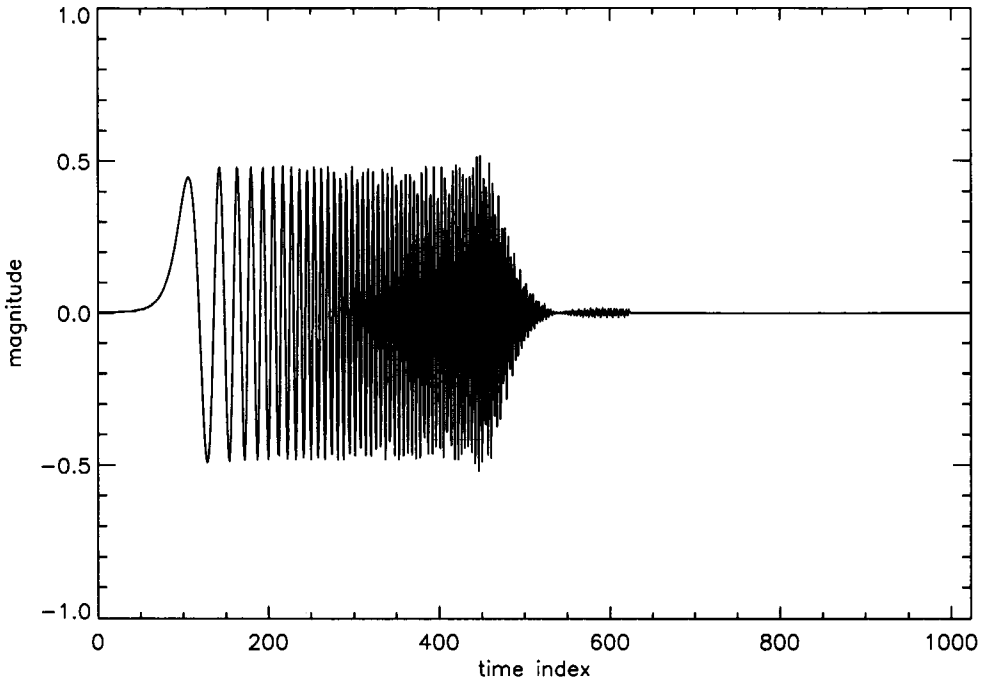


Figure 10.2.F7

After application of this filter the impulse response function of concatenated channel and cancellation filters should look like a bandlimited delta function. Figure 10.2.F7 shows that this is indeed the case.



The Restored GCR Signal from the Cascaded FIR and IIR Filters (Echoes are Removed)

Figure 10.2.F8

The corresponding shape of the GCR should be identical to the original GCR of course. This is substantiated by Figure 10.2.F8.

CHAPTER 11

IMPLEMENTATION OF THE HARDWARE

In this chapter we will describe the hardware aspects of the system that was implemented for the later phases of the field tests in the U.S. It also served as a prototype for a professional product that has been made commercially available and was up to now purchased by various TV stations and also by cable operators (for use in their head-ends). This product has also been adapted for use in field tests in other countries with different TV systems.

11.1 The Overall Structure of the System Hardware

Basically the hardware structure consists of three blocks as illustrated in Fig. 11.1.F1:

1. An input block which receives the analog baseband composite video signal, performs the analog preprocessing that is required in the context of echo cancellation and then converts the signal to digital form.

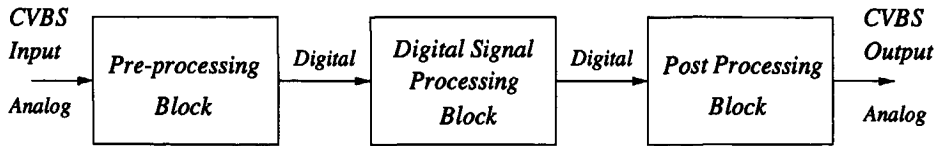


Figure 11.1.F1

The Major Blocks of the Overall System

2. A digital processing block that takes care of all the tasks that are necessary for echo cancellation, such as channel characterization by cross correlation, calculation of the filter coefficients of the FIR and IIR filter by adaptation and convolution (see chapter 9), the actual use of these filters for cancellation as well as control of e.g. the magnitude of the steps in the adaptation process and switching off the cancellation in case of malfunctioning.

3. An output block for D/A conversion, anti-aliasing filtering and amplification.

11.2 The Input Block

This block, see Fig. 11.2.F1, receives the echoed analog baseband composite video signal, which is already adjusted to the standard [1] voltage range of -40 IRE to 100 IRE, through a capacitor that serves for dc separation. This creates the possibility of random drift after this capacitor. The first operation to be performed is to correct for this drift by clamping. The clamping process has to be slightly

different from its counterpart in a conventional receiver [2] because one of the main reasons for non-random variation of the dc level is the occurrence of echoes. If the usual "hard" clamping by means of a diode is used we essentially reduce not only the random drift but also the drift that is caused by echoing. This means that hard clamping with a diode only, would constitute a partial cancellation of the effect of echoing by means of a non-linear process. This introduces non-linear deformation of the signal that can never be corrected by the linear echo cancellation process as described in chapter 8. The nonlinearity can be reduced by "softening" the clamping through the introduction of a resistor in series with the diode at the price of slowing the clamping down over several line durations.

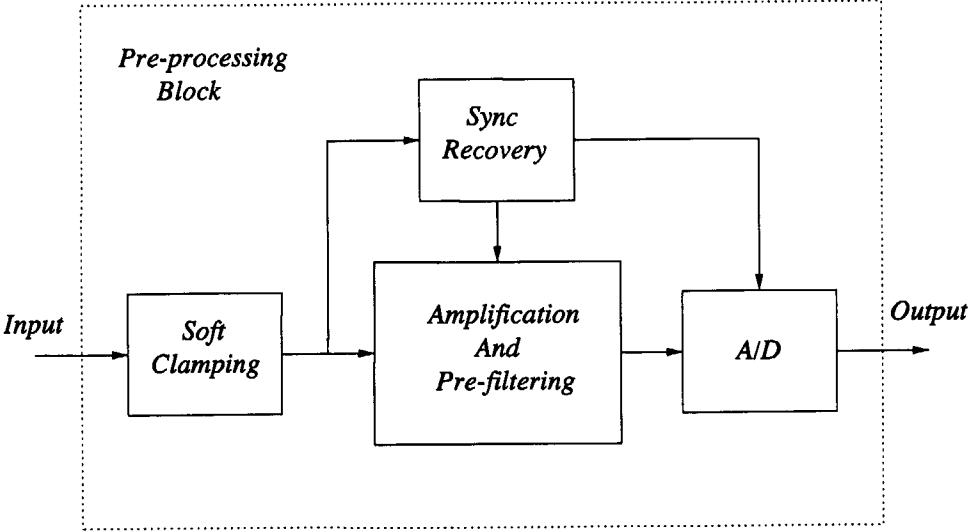


Figure 11.2.F1
The Input Block of the System

Subsequently the horizontal and vertical synchronization pulses, which are contaminated by echoing and therefore can be missed by conventional synchronization procedures, are replaced by standard synchronization pulses, generated locally and timed by a proper local lock-in mechanism that remains effective if a number of the original synchronization pulses are not properly detected. This then makes it possible to synchronize the key-in periods of the phase locked loop (at 8 times color sub-carrier frequency) that has to provide the clocks in the system. The only hardware element that actually runs at this high frequency is the microprocessor. Everything else runs at the derivative frequency of 4 times sub-carrier.

After this procedure A/D conversion is done by means of a 10 bit A/D converter. Of the ten bits it delivers only 9 are used. This is one more than in conventional digital video processing, which is necessary because echoing increases the dynamic range of the video signal.

11.3 The Digital Processing Block

The digital processing block, see Figure 11.3.F1, essentially consists of the μ -processor and the programmable cancellation filter. The μ -processor performs all the digital processing tasks except the actual filtering. It is a Texas Instruments TMS320C25 device that uses 8k words (16 bits per word) fast memory, used for Direct Memory Access (DMA) in grabbing the incoming echoed GCR. Furthermore its memory has 96k 16 bit words of RAM as well as 32k 16 bit words of EPROM. Its programs have been written in C language, compiled into the device's machine code and burned into the EPROM.

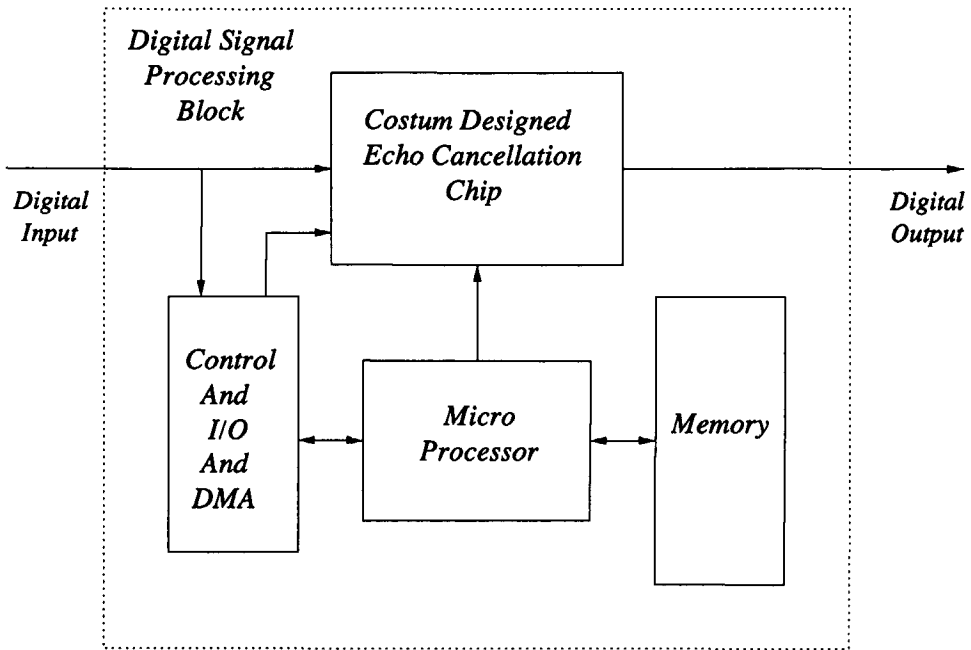


Figure 11.3.F1
The Digital Processing Block

One additional piece of hardware is "watch dog" circuitry that resets the system in case of any malfunctioning in the μ -processor. This watch dog expects, regularly in the course of time, a signal from the processor that the latter is able to run its programs without deadlock or interference from electrical glitches, hardware and software failures and the like. Obviously the processor has been programmed to provide or not to provide these "all is well" signals. If it fails to send such a signal the watch dog causes the processor to reboot and start again and at the same time has the video signal bypass the echo cancellation filter, so that echo cancellation stops. As a result the malfunctioning in the processor never makes the system

worse than a conventional receiver. If after booting the processor performs as it should it will send the "all is well" signals again and the bypass accordingly is lifted.

We note that apart from the undesirable situation described above there is the danger that instabilities and other irregularities occur in the convergence of the adaptation process (see 8.2) for the FIR filter, in the IIR filter, which after all contains a feedback loop, or because of a notch in the channel characteristic. In all of these cases the processor has been programmed to take counter measures. Therefore we defer a discussion of these circumstances to the next chapter, where we will describe the software of the system.

The actual programmable cancellation filter is a custom Integrated Circuit. The custom design was mainly done by our team, which contracted TLW Inc. of Boston to do the low level design. The chip was manufactured by Hewlett Packard. The system contains two of these chips, each containing 9 clusters of 180 taps with 8 bit by 8 bit multipliers. The delay between the clusters can be chosen to vary between wide limits (each cluster can be placed at a delay from -3 μ sec to 45 μ sec) in order to take advantage of the fact that often echoes occur in the course of time in groups. If on the basis of the channel characterization, the clusters are placed along the time axis in accordance with these groups, the total echo delay that can be handled by the system can be maximized at reduced cost in terms of (expensive) multipliers [3].

A proprietary feature of the IC is the possibility to move the clusters between the FIR and the IIR portion of the filter in an elegant manner [4].

11.4 The Output Block

The output block, see Figure 11.4.F1, receives the echo free digital video signal from the output of the cancellation filter. It converts the signal to analog form for which of course an analog

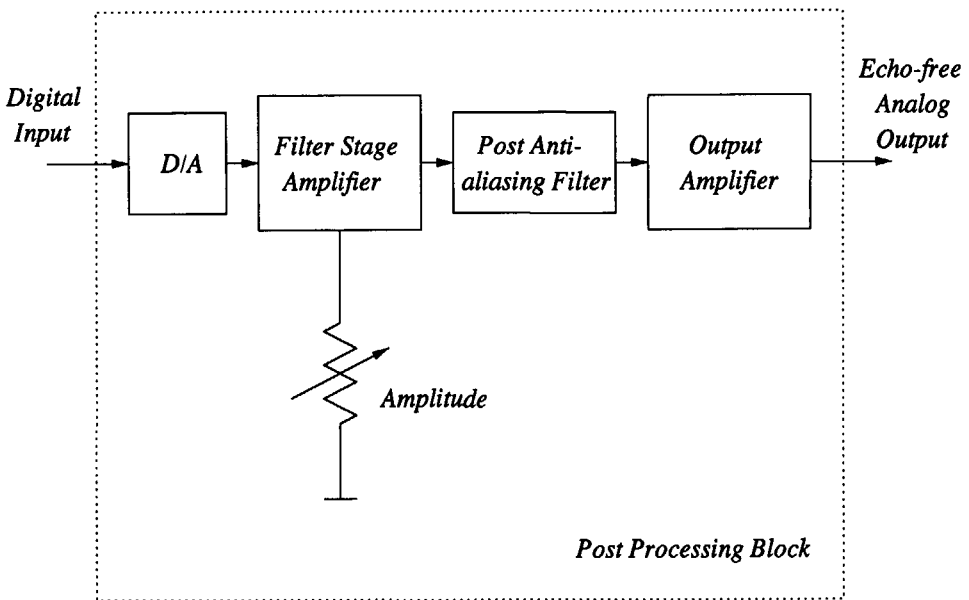


Figure 11.4.F1
The Output Block of the System

anti-aliasing filter has to be included. It also includes two wide band amplifiers, one in the anti-aliasing stage and the other in the final output stage of the system.

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CHAPTER 12

IMPLEMENTATION OF THE SOFTWARE

The software for the system, written in C and compiled into the machine code of the μ -processor is organized in such a way that at the highest level the same sequential series of instructions is executed over and over again. Inside this sequence there are various loops that may be executed a number of times before the program moves on to the next step of the main sequence. Also, on the basis of intermediate results, the program may have to reset parameter values that govern its control and even go back to a previous phase of the sequence to adequately use these new parameter values.

The existence of the basic sequence allows us to describe, in the following sections, the program to first order in terms of subsequent computational events. At certain points in this description we will indicate where and why setting a new parameter value and/or jumping back to a previous stage is necessary.

The detail into which we go in the following is going to be restricted in view of the size of the total program. We hope however to be successful in indicating with sufficient clarity what the essential features of the program are.

12.1 Capture of the GCR

As noted in 1.3.3 the GCR is located in unechoed form on line 19 of each field and in echoed form may also occupy portions of line 20 while echoing may cause line 18 to leak into line 19 and vice versa. Furthermore (see 1.3.2) the 8 field cycling period that is used for sending the original and capturing the echoed GCR removes the color burst, sync pulse and all non-varying information and averages all varying information on these lines which results in a capture process that is usually sufficiently reliable.

From this it will be clear that the program starts with performing the 8 field cycle capture. In fact the commercial professional cancelers that were produced use a factory setting for 32 cycles of 8 fields in order to improve, at the cost of losing speed, the signal to noise ratio of the process, while up to 64 cycles of 8 fields can be set by the user.

As described in 11.3 this capture uses the fast DMA memory each time it fetches a particular GCR. After each capture of a GCR a full field period is available to place the result in RAM. The latter process has been made additive during each particular cycle, in order to perform the averaging.

After this averaging the result is investigated according to certain criteria in order to establish whether indeed a GCR has been captured and if so whether it has not been distorted too much for echo cancellation to be helpful. This obviously is necessary in particular because not all TV stations transmit the GCR as yet: "cancellation" on the basis of a non-GCR on line 19 would result in a disastrous picture quality.

If the result of this test is negative the program goes back to the capture phase again. This process repeats until a "legal" GCR is detected.

The above has been summarized in Fig 12.1.F1

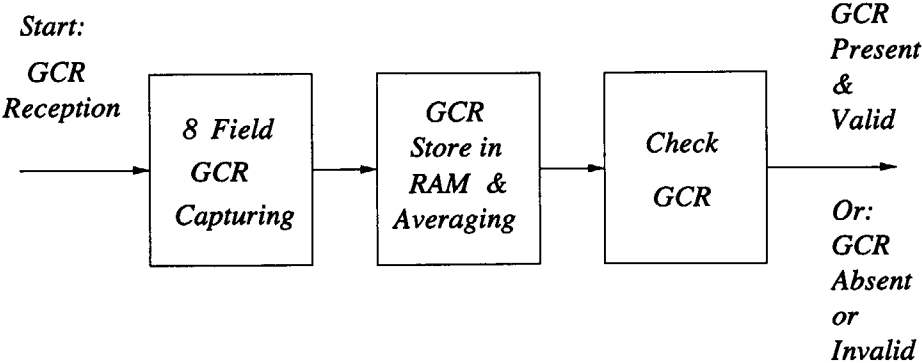


figure 12.1.F1
The GCR Capturing Process

12.2 Channel Characterization

The next step is performing channel characterization by cross-correlating the received GCR with the original GCR as stored already in memory. This correlation procedure is done by using a Fast Fourier Transform (FFT) [1][2][3] of the former and using the complex conjugate of the FFT of the latter. The results are multiplied for each frequency and the product is converted back to the time domain by an inverse FFT. This procedure relies on the well known three facts [4] that (1) correlating two functions is equivalent to convolving one of them with a time reversed version of the other, that (2) convolution in the time domain is

equivalent to multiplication in the frequency domain and that (3) the complex conjugate of a Fourier Transform of a real function of time is the Fourier transform of the time reversed version of that function. The procedure is much more efficient than actually correlating in the time domain because multiplying is a much cheaper operation than convolving.

The above paragraph has been summarized in Fig 12.2.F1

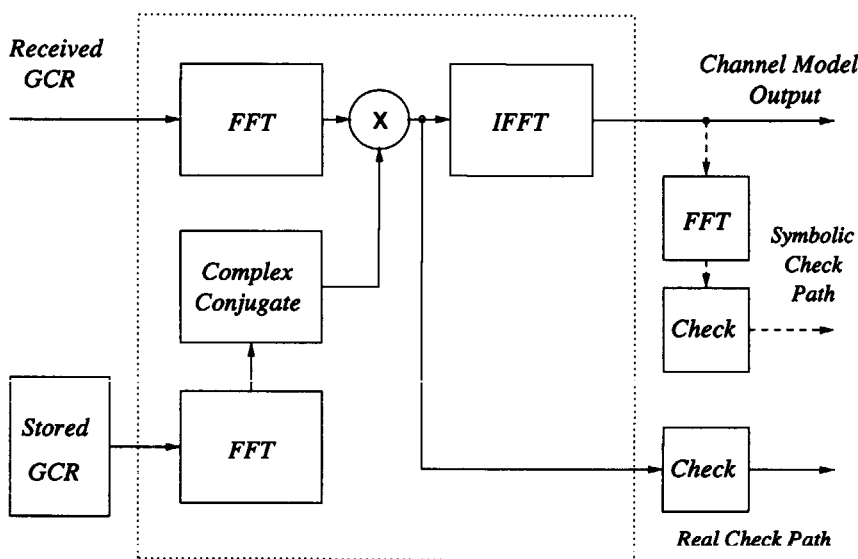


Figure 12.2.F1

The FFT-based Cross-correlation Process

The particular FFT procedure that was used was tailored to the needs of the (fixed point) μ -processor and the GCR amplitude in order to ensure the required bit

resolution and dynamic range.

We recall here from 4.2.2 that the cross-correlation that we have thus obtained is identical to the impulse response function of the channel due to the flat frequency distribution of our GCR and can therefore serve to orchestrate the next phase: actual echo cancellation.

Before we pass on to that phase of the program we should remark that on the basis of the channel model that we obtained, there may be reasons here to reset parameter values in the program.

In order to investigate whether this is necessary the program checks the frequency domain channel model before the inverse Fourier transform (IFFT) operation (Fig.12.2.F1). This is of course the frequency characteristic of the channel and is equivalent to the output of the symbolic check path, which may display one or more high peaks and/or deep notches.

If peaks are found the value of μ in the adaptation process for finding the coefficients of the FIR filter (see 9.1.E5) must be reset. This can most easily be seen by means of a Fourier Transform of 9.1.E5:

$$C_{k+1}(\omega) = C_k(\omega) + 2 \mu E_k(\omega) H_k(\omega) \quad (12.2.E1)$$

We see here that in case of a peak, where $H(\omega)$ is large, the approximation steps will tend to be relatively large which would endanger the convergence of the process as explained in connection with 8.4.E7, which is the equation correspond-

ing to 9.1.E5 for the case of channel characterization by adaptation. This threat to the convergence can be counteracted by resetting μ to a smaller value.

We also see that in case of a notch, where $H(\omega)$ is small or possibly zero, the approximation steps will tend to be relatively small or zero. Hence there will be a slow down of the convergence process or even a reduction of its speed to zero. This could be counteracted by extending the number of approximation steps (default value 16). We have chosen not to do so because complete cancellation of a notch obviously will result in a boost of the channel's frequency response at the frequency concerned and a corresponding boost in the video noise at that frequency. In fact we have experienced that a disastrous "blow up" of the picture can occur when there is a deep notch. Limiting the number of approximation steps makes the cancellation incomplete and thus reduces the boost and the danger of a blow up.

12.3 Calculation of the Coefficients of the Cancellation Filters

This section starts out with an explanation on the distribution of resources between the FIR and the FIR1 filter in the IIR loop. After that we enter into a discussion of the calculation of all the filter coefficients. This discussion as a whole is illustrated in the following Fig 12.3.F1

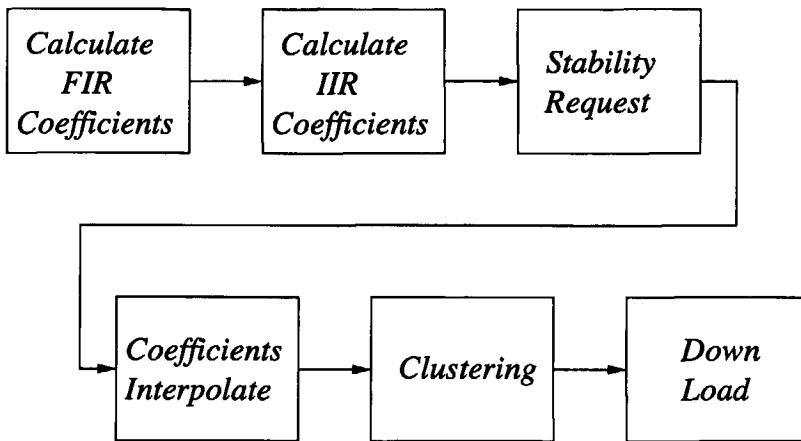


Figure 12.3.F1

The Procedure for Calculating the Coefficients of the Cancellation Filters

12.3.1 Distribution of Taps Between FIR and IIR Filters

As has been said in 11.3 the two identical custom ICs that each contain 9 dynamically locatable clusters of 20 taps among other things offer the possibility to distribute these clusters at will between the FIR filter and the IIR filter. On the basis of our experience and the considerations to be presented in the next paragraph we nevertheless let the software initially assume that it has a fixed number of 200 taps at its disposal for the calculation of the FIR filter. As a consequence the software of the actual system that we are describing does not have provisions to switch clusters between FIR and IIR.

As we have argued before (2.3.3.2) on the one hand the FIR can be relatively short because pre-echoes in practice tend to precede the main signal by only some 5 μsecs , but on the other hand its actual length has to exceed these 5 μsecs because of the property of FIR filters that they cancel echoes at the cost (see 2.3.3.2) of creating smaller new echoes at double the (in this case negative) delay time. These echoes have to be removed also and in the process new ones of smaller amplitude are created again and so on. Furthermore the FIR is used to cancel as well some nearby post-echoes. It will thus be clear that the FIR filter has to cover the time span from about $-10 \mu\text{secs}$ to about $+2 \mu\text{secs}$, which in view of the sampling frequency of 4 times color sub-carrier frequency would require the 200 taps mentioned above.

The allocation of resources to the FIR filter is this generous because the IIR filter's coefficients will be calculated on the basis of those of the FIR filter. Consequently shortcomings in the FIR filter will propagate into the IIR filter and should be avoided as much as possible. Also the calculation of the FIR filter involves convolution of error vector and impulse response of the channel for each approximation step, in which each next generation coefficient is influenced by all present generation coefficients. This suggests that during the calculation more precision is needed than in the final result. We found this empirically to be the case by seeing in practice that in the final result the echoes in the "earliest" 50 taps practically vanish. Consequently we need only 160 taps or 8 clusters for the FIR filter in the hardware, leaving 200 taps or 10 clusters available for the IIR.

As we will see below, once the ideal coefficient values for the IIR filter are known, we will distribute these (still) scarce resources in such a way that their use for the IIR is as effective as possible.

12.3.2 Adaptation of the Coefficients of the FIR

The adaptation of the FIR coefficients is done according to the Least Mean Square algorithm [5] described in 9.1. The number of approximation steps has been set to the constant value of 16 on the basis of analysis, simulation and experience, taking into account the considerations on notches in the frequency characteristic of the channel as presented in 12.2. After completion of these 16 approximation steps the resulting coefficient values are committed to RAM in preparation for further processing as described in 12.4 and for calculating the coefficients of the IIR filter.

12.3.3 Calculation of the Coefficients of the IIR Filter

The calculation of the coefficients of the IIR filter is done as described in 9.2. The convolution of the impulse response of the channel and the FIR filter that has to be performed is done in a way that is similar to the way in which we did the cross-correlation of received and original GCR, using FFT and multiplication followed by inverse FFT: both vectors are Fourier transformed, multiplied and the product vector is converted back to the time domain after subtracting the band-limited and time windowed $\delta(t)$ as described at the end of 9.2, which is depicted by Figure 12.3.3.F1. The resulting new coefficient values for the IIR filter are stored in RAM for further processing as described in 12.5.

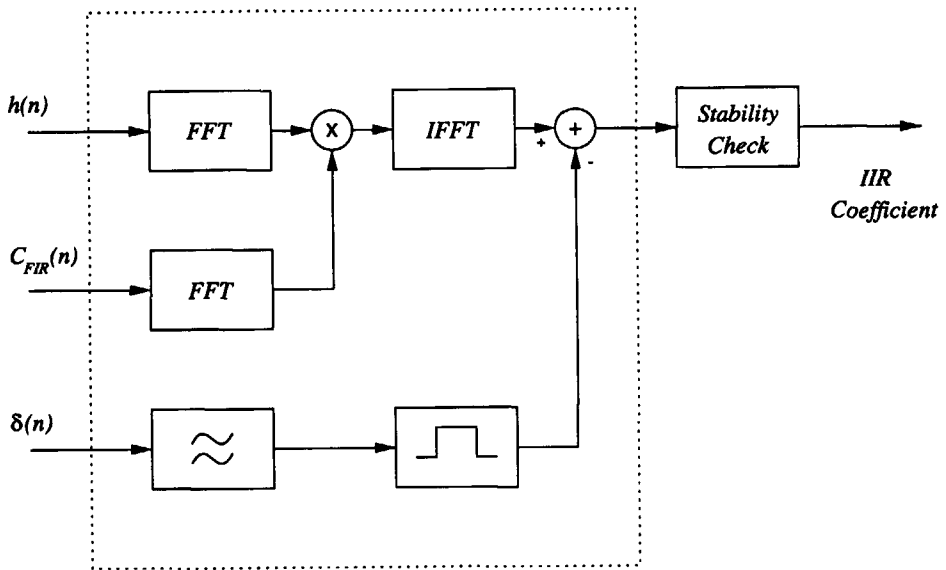


Figure 12.3.3.F1

The Calculation Process for the IIR (FIR1) Coefficients

Obviously, since the IIR filter can be described in the z domain as having the characteristic $\frac{1}{(1 + H(z))}$, where $H(z)$ is the convolution of the real channel and the FIR filter minus the band limited and time windowed version of the $\delta(t)$ function in 9.2.E2, it becomes unstable when $H(z) = -1$ for some z value outside the unit circle in the z plane [6][7]. If we write

$$H(z) = a(0) + a(1)z^{-1} + \dots + a(n)z^{-n} \quad (12.3.3.E1)$$

then

$$\left| (H(z)) \right| = \left| (a(0) + \dots + a(n)z^{-n}) \right| \leq \left| a(0) \right| + \dots + \left| a(n)z^{-n} \right| .$$

Outside the unit circle this expression is smaller than

$$\left| a(0) \right| + \dots + \left| a(n) \right| \tag{12.3.3.E2}$$

Therefore if the latter expression is smaller than 1, the absolute value of $H(z)$ is smaller than one and thus $H(z)$ cannot equal -1 and the IIR filter is guaranteed to be stable.

We thus certainly avoid instability if the value of 12.3.3.E2 is constrained to be less than 1. In order to achieve this we have chosen to let the software reduce all coefficients by 20% as many consecutive times as necessary until 12.3.3.E2 does not exceed 1. Of course this results in incomplete cancellation but the result was seen to be better than no cancellation at all in the rare cases where this correction was necessary.

12.4 Coefficient Interpolation

The filter coefficients cannot be downloaded directly into the cancellation filter since a too discontinuous change in the coefficients can cause perceptually unacceptable discontinuities in the appearance of the picture. To prevent this it turned out to be necessary to interpolate 7 intermediate steps, to be computed on a linear basis. The result of this interpolation again is stored in memory.

12.5 Clustering of the Calculated Interpolated IIR Coefficients

The number of available IIR taps in the hardware is exceeded by far by the number of tap samples that are calculated for the IIR. For that reason the 10 available clusters still available after allocating 8 clusters to the FIR filter will have to be distributed in an intelligent way. We have chosen to do this by letting the software find, for each calculated and interpolated set of coefficients, the largest coefficients and to allocate clusters to these, priority corresponding to coefficient magnitude. For each cluster its position in the course of time relative to that coefficient is then determined in such a way that, again with priority for the largest nearby coefficients, as many large coefficients as possible are covered by the cluster. This is done until all the available clusters have been used.

12.6 Downloading the Filter Coefficients into the Cancellation Filter

The clustering process described in the last section is of a combinatorial nature. Therefore trial and error is involved, especially if two of the largest coefficients are located close to each other. This is a time consuming process. It turned out that in our system the time required was considerably longer than a field period. For that reason the software was so constructed that for each interpolation the coefficients are downloaded into the cancellation filter as soon as possible after they become available.

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CHAPTER 13

POTENTIAL REASONS FOR FAILURE OF THE SYSTEM

At this point our description of the system is complete. In this description we have occasionally mentioned reasons for possible failure but we have not yet collected them in a usable list. Before we turn to the tests to which the system has been subjected we present such a list in the present chapter. It also will have to serve as an explanation of the fact that the test results, though significantly better than for the competing systems, were not so perfect that the restored TV programs had studio quality.

The list will be presented below in the form of a series of short sections.

13.1 Noise

Noise affects channel characterization as well as the cancellation itself.

In the channel characterization process we deliberately chose the correlation method since it is relatively insensitive to noise as compared with frequency division and still simple as compared with adaptation.

In cancellation there is not very much that one can do about noise in general with the exception of the case where the channel characteristic has a deep notch in the frequency domain. An attempt to fully cancel echos will boost the noise around the frequency where the notch occurs, sometimes to the extent that instability sets in. An obvious way to counteract this is to avoid complete cancellation in this case as described in 12.2.

13.2 Time Varying Echoes

Airplanes, trees and antennas moving by wind etc. can cause the echo pattern to change too quickly for the cancellation system to cope with. As explained in chapter 2 in NTSC M we need at least 8 fields of data gathering in order to cancel the influence of constant information as well as of the color burst. This takes about 150 milliseconds, long enough for the human perceptual system to see the image to some extent. Accordingly we cannot expect to be able to satisfactorily cancel echos that vary substantially in 150 milliseconds. If we average over periods longer than 8 fields in order to get rid of time varying information on lines 18 and 20, the system becomes still slower.

13.3 Deep Notches in the Spectral Response of the Channel

The notches have been touched upon already. Suffice it to state here that during many field tests we were confronted with this phenomenon. As has already been pointed out we coped with this in our system by giving up on complete can-

cellation: it is better to have only partial cancellation than to be confronted with instabilities or excessively noisy phenomena on the screen.

13.4 Instability in the IIR Filter's Loop

We have paid attention to this situation in 12.3.3. It has been explained there that we took recourse in this case to a method that prefers to accept false alarms to non detection of instability. In fact we saw that the software in case of potential instability reduces all coefficients in the IIR so that, again, incomplete cancellation results.

13.5 Non Linearities

We have stated repeatedly that we purposely did not attempt to attack non linear phenomena, the reason being that echos constitute a linear distortion and can be canceled by linear and therefore relatively simple methods. However when non linear distortion occurs for whatever reason, echo cancellation methods cannot help and a non perfect picture will result that will qualify correspondingly poorly in tests, especially when subjective evaluation by naive observers is used. As stated in chapter 2 one source of non linearity is the use of a-synchronous or quasi synchronous detection. Our whole treatment of echo cancellation therefore explicitly assumes synchronous detection in the receiver.

13.6 Echo(s) and Main Signal of Comparable Magnitude

If the echo pattern is so strong that it is comparable in magnitude to the main signal, various choices that the system has to make become difficult and the system may perform poorly in that situation. In spontaneous situations and under normal conditions during the tests we have not observed this phenomenon but it can be evoked by manipulation of the antenna. Consumers however will try to have their antenna directed in the most optimal direction and this will be sufficient to avoid the phenomenon.

13.7 Hardware and Software Limitations

There are always trade-offs that have to be made and systems can always be made better at a higher price. This applies to hardware and software and we have mentioned numerous trade-offs that we made in the text: length of filters, accuracy of timing, bit allocations etc.

Here we would like to draw attention in particular to our clustering of coefficients as mentioned in 12.5. Clustering provides efficiency in the use of available filter taps, but not in all cases: if the echos are very vague, smeared as it is called, one simply needs many more coefficients or taps than we were able to use and the quality of the cancellation decreases accordingly. We have seen this happen in particular field situations and some of the tests were explicitly intended to compare various systems on their "smeared echo" performance.

CHAPTER 14

TESTS AND CONCLUSIONS ON COMPETING SYSTEMS IN THE U.S.A. (1992)

Two professional organizations have organized extensive comparative tests on proposed systems for echo cancellation on a national level: The Broadcast Television Association (BTA) of Japan in 1989 and the Advanced Television Systems Committee (ATSC) of the USA in 1992. The latter benefited from the help of some of its member organizations: Association for Maximum Service Television, Cable Television Laboratories, Electronic Industries Association, National Association of Broadcasters and Public Broadcasting Service. In both cases analog GCRs as well as Pseudo Random Sequences were considered and in both cases an analog GCR prevailed.

14.1 Tests in Japan in 1989

BTA's work in Japan preceded that of the ATSC in the USA. It led to a GCR standard in Japan that is based on the use of an integrated bandlimited δ -pulse as a GCR.

Without going into too much detail a short explanation of this may be given here. Obviously a bandlimited δ -pulse has a flat spectrum within the bandwidth of the channel, which is a required property for a GCR, as discussed in chapter 4. On the other hand the energy content of a band limited δ -pulse is poor, due to the short duration of its main peak, which makes the signal noise sensitive. For this reason the Japanese system integrates the δ -pulse to a "wavy" step-function prior to transmission [1]. This step-function has an infinitely long duration, which must be cut off at a certain moment. In any case the energy content of this signal is relatively large. The process does not affect the potential for echo cancellation as can be easily seen in the frequency domain: integration means division by $j\omega$ in the frequency domain so that after passing through the echoing channel we have the product of $\frac{1}{j\omega}$, the frequency characteristic of the band-limited delta function and the frequency characteristic of the channel. The first two factors are known and can be divided out, which leaves us with the frequency characteristic of the channel. It should be noted however that at zero frequency we have to cope with the very large magnitude of the integrated signal (which is only finite due to the finite time interval over which the integration is done), that at high frequencies, where the color information of the video signal is concentrated, the GCR is relatively weak in energy, and that the cut-off of the step-function causes difficulties for post echoes with long delays.

14.2 Tests in the U.S.A. in 1992

The work of ATSC took place somewhat later and was initially organized on the basis of a contest between the five qualified proponents by means of a field test program for broadcast (five proponents, three channels and about a hundred loca-

tions, which amounts to about 1500 separate tests) [2] and a test program for cable using actual cable systems as well as laboratory tests [3].

On the basis of our previous considerations we can restrict ourselves here to the following brief descriptions of the competing systems:

1. AT&T/Zenith Electronics Corporation [4]: based on a Pseudo Random Sequence.[5]
2. BTA [6]: as described above, based on an integrated band-limited δ -function.
3. David Sarnoff Research Center/Thomson Consumer Electronics [7]: based on a Pseudo Random Sequence.
4. Philips Laboratories [8]: the system described in this thesis.
5. Samsung Electronics [9]: based on an extension of the Pseudo Random Sequence concept.

These systems were all fully tested as described above. For the Broadcast tests subjective evaluation was used on the basis of the CCIR Quality and Impairment scales [10]. An extensive report [2] was prepared which had as final conclusion that our system "consistently outperformed the other four systems".

It might be expected by the reader that we quote more specific test results in the form of tables or graphs from reports on tests. We have chosen not to do so in this thesis because we found it virtually impossible to make a responsible choice

from the abundant material in the reports. They should either be consulted in full (as we did) or not be consulted at all.

14.3 Subjective Assessment and its Results in the U.S.A. Tests

In this section we comment on some general aspects of evaluating the performance of systems like the one described in this thesis (14.3.1) and then quote the overall results of the subjective tests in the USA (14.3.2). We finalize our account of the U.S.A. tests in the next section (14.4).

14.3.1 Evaluating System Performance

Up to this point we have refrained from making any comments on the actual improvement in picture quality that the described system produced. There are various reasons for this, mentioned in the following list.

1. Showing or quoting specific instances of picture improvement, however impressive they may be, is not convincing in itself without detailed information on the circumstances under which the result was obtained, how normal and abnormal these circumstances were, how the result compares with those of other systems, what kind of equipment was used, what the settings of the equipment were and so on.
2. The system that we created of course is the result of considerable theoretical consideration, experiments in the field and trade-offs in both

hardware and software. The theory should be correct of course but strictly speaking has to be tested experimentally on its own, which can only be done with a perfect implementation, which would be prohibitively expensive. Consequently in practical tests one cannot separate the quality of the GCR as such from the necessary shortcomings of the implemented system. A judgement on the theory alone can of course be obtained to some extent from simulations as presented in section 10.2, though also here GCR quality and method of cancellation cannot be separated.

3. In fact for the same reasons one cannot adequately compare the various GCRs that have been proposed recently or in the past.

What has been done during the tests mentioned in the previous chapter seems to be about the only sensible way to come to an acceptable decision as to system choice: compare subjective picture quality under a large variety of circumstances (channel, location, identity of the observer), somehow average the results and complement this with objective laboratory measurements, followed by a voting process in which a broad variety of TV related parties participate. Obviously this process is not perfect. The specific interests of the various participants will influence the result but hopefully these influences cancel out to a large extent if the number of parties is as large as it actually was in the process under discussion.

14.3.2 Results of subjective tests in the USA

We accordingly think that the most responsible and satisfying way to describe the result of our efforts is, apart from quoting simulations, to quote the overall

result of the subjective tests for ATSC. This result was formulated at the time [11] in terms of improvements on the CCIR impairment scale [10] produced by the systems under test under the wide variety of circumstances that was used. This scale reads as follows:

5. Imperceptible,
4. Perceptible but not annoying,
3. Slightly annoying,
2. Annoying,
1. Very annoying.

The average impairment before cancellation was determined to be 2.6, in other words before cancellation the average picture was considered to be between "annoying" and "slightly annoying". The improvements produced by the systems under test was judged to be

BTA	1.137
Samsung	0.233
AT&T/Zenith	0.030
Philips	1.517
Sarnoff/Thomson	0.880

In terms of average CCIR impairment this means that our system achieved $2.6 + 1.5 = 4.1$, thus raising the subjective impairment level to better than "percep-

tible but not annoying". All other tested systems showed less improvement. It should be noted here that according to the expectations described in chapter 7 the systems with a purely analog GCR performed best.

Also, as a concluding remark we would like to point out again that although good echo cancellation is expected to remove not only echoes but in addition all linear distortions from the video signal, it will not remove other impairments such as noise and non linear defects. Therefore we cannot expect, even perfect, echo cancellation to raise the CCIR impairment level to 5 under non ideal conditions.

The tests for cable had a less clear outcome [3]: it was felt that our system was not clearly better than the Sarnoff/Thomson system.

14.4 Final Series of USA Tests

As a result a new series of tests was organized between the Philips and Sarnoff/Thomson systems in original form and a combined system in two versions, created in collaboration between David Sarnoff Research Center and our Laboratory. It used Pseudo Random Sequences as well as our proposed GCR in two versions.

As to the broadcast tests it was concluded [11] that "the Philips original system consistently exhibited the highest performance".

The conclusion of Cable Television Laboratories was [12] that for cable application our hardware with our original GCR performed slightly better than with

the best of the two combined signal versions and that the Sarnoff hardware gave the best results with the same combined signal. Here it should be noted that tests in which the hardware of either proponent used the original GCR of the other proponent could not be conducted.

The Cable Labs report finally stated that, and here we quote:

- a. The Philips hardware did better on long ghosts and smear ghosts.
- b. Philips did better on time-varying long ghosts.
- c. Sarnoff did better on time varying short ghosts.
- d. Both proponent's ability to track periodic ghosts required minimum periods in the tens of seconds duration. Neither system demonstrated complete cancellation.

14.5 Conclusions

In subsequent meetings of ATSC committees at three levels of increasing authority the result of the voting was that our system should be recommended to the FCC for promotion to a mandatory standard for the USA. There were no votes against this. A few members abstained. This led eventually to a "Report and Order" by the FCC [13], dated May 1993, in which the FCC dictated that "line 19 of the VBI should be reserved for the optional but exclusive use of the Philips GCR signal".

In September 1995 both our laboratory and the present author were awarded an "Emmy Award" for our work on echo cancellation by the Academy of Television Arts & Sciences.

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† RFP: Request For Proposal.

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† RFP: Request For Proposal.

CHAPTER 15

INTERNATIONAL ASPECTS

The many different TV systems in the world require slightly different echo cancellation systems, that have to be adopted in some formal way, depending on the country or group of countries in question. For that reason we have to conclude this thesis with a chapter on international aspects. It will consist of two parts: a technical part to indicate what the technical issues are that have to be considered for adaptation of our system to local requirements and a (short) part on formal matters regarding international standardization.

15.1 Technical Considerations

Without going into detail we list below the various technical issues that we had to consider before actually adapting the hardware and software of our system for field tests in various countries such as Argentina, Australia, Austria, Brazil, Canada, China, France, Germany, Mexico, New Zealand, United Kingdom and Singapore.

1. In the first place there is channel bandwidth which may vary about one Mhz

between systems. As we have stated many times in previous chapters the GCR has to cover the whole channel bandwidth, which in our case can be realized independently as discussed in chapter 5.

2. Since different systems have different color subcarrier frequencies, the phase locked loop that governs the clock system has to be adjusted.
3. In connection with that the method of locking to the color burst has to be redesigned for PAL systems, since in those systems the phase of the color burst changes by 90 degrees every field [1] instead of the 180 degrees that we know from American NTSC.
4. The 8 field cycling period for transmitting and recovering the GCR will not work for all TV systems and has to be reconsidered.
5. The duration of the GCR has to be changed in order to optimize system performance in view of the contents of neighboring lines of the VBI. This was done in countries which use Teletext [2][3] in order to avoid long averaging of this time varying signal: in PAL we use a GCR duration of only 23.3 μ sec in order to be able to combat post-echoes with a delay of 40 μ sec without interference from teletext on the next line.
6. In SECAM there is no color burst [1] and the clock has to lock to the horizontal synchronization pulses.
7. The pedestal/amplitude combination of the GCR (see 1.3.1) must be optimized in accordance with local regulations.

It will be clear from chapter 5 that the adjustable parameters of our GCR fully allow all adjustments to the GCR that the above points require. It will also be clear that adjustments to the GCR alone are not sufficient: both hardware and software need to be modified too in order to adapt to various TV systems.

Our team has realized these modifications for field tests in all of the countries mentioned above. The field tests were universally successful which led to the international standardization as described in the next section.

15.2 Standardization Aspects

The International Telecommunication Union has issued a document on International Standardization of GCRs [4] dated April 1994 stating that for conventional TV systems (NTSC, PAL, SECAM) a broadcaster wishing to integrate a GCR signal should use a GCR signal complying with the specifications given in Annex 1 of the document for three systems. Two of these are respectively described as a standard already adopted in Japan and one to be adopted in Korea. The third is described as an ATSC standard (see chapter 11) for NTSC in the USA, to be adopted in Canada, and under evaluation for PAL and SECAM-625 line systems in Europe, Australia and New Zealand.

The Joint Technical Committee of the European Broadcasting Union, the Comité Européen de Normalisation Electrotechnique and the European Telecommunications Standards Institute submitted a document [5] for public enquiry in 1996 containing a proposed standard for a GCR in 625 line TV systems, to be

placed on line 318, that strictly adheres to our proposal for a family of GCRs. As mentioned in point 5 of the previous section the duration of this GCR was fixed at 23.3 μ sec in connection with the variable contents of the next line, such as Teletext.

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CHAPTER 16

PROSPECTS FOR FURTHER WORK

Prospects for further work necessarily constitute a speculative subject. Nevertheless we venture the following suggestions:

1. Consumer echo cancelers have been fabricated and have been sold to consumers for a few hundred dollars. Their performance is reasonable but can be improved. This clearly is a subject worth of further study.

2. Application to digital communications of various kinds is an obvious next step if one realizes that even if the content of transmissions is digital, channels are always analog. Of course comparisons with channel equalization as practiced in the field of telephony are necessary.

3. Applications in Radar, Sonar and Seismology should be considered. They will not have to do with cancellation of echos but with analysis of echos: characterization rather than cancellation.

4. A GCR like the one we have proposed may be useful as an excitation signal for mechanical diagnosis.

5. It is intriguing to speculate about extension to more dimensions: is it for example possible to correct the image produced by a large but faulty mirror by methods similar to those described in this thesis?

CHAPTER 17

CONCLUSIONS

The conclusions that can be drawn from the work reported in this thesis are the following:

1. The creation process of echos in television is linear when viewed from base band on the transmission side to base band on the reception side.
2. Consequently echo cancellation can be a linear process as well.
3. Good echo cancellation can restore all linear defects of the received signal.
4. Linear echo cancellation cannot be expected to restore non linear defects.
5. It is possible to list a set of requirements for a Ghost Cancellation Reference (GCR) signal that is rich enough to allow a straightforward design of an optimal family of GCRs and not so restrictive that it prohibits a sensible design.
6. A flat spectral distribution in a GCR permits the use of an efficient and

noise robust method of channel characterization by cross correlation of the distorted received GCR with the original undistorted GCR.

7. Cancellation of echos by means of programmable digital filters is theoretically possible, in the sense that all the mathematics for this is available, as well as practically feasible, in the sense that with present day technology and processing power satisfactory cancellation can be obtained at affordable cost.
8. Pseudo Random Digital Sequences are less than optimal for use as GCR.

APPENDIX

**INFORMATION AND TECHNICAL SPECIFICATIONS OF
PHILIPS GB-180 GHOSTBUSTER CHIP**

Philips GB-180 Ghostbuster Chip

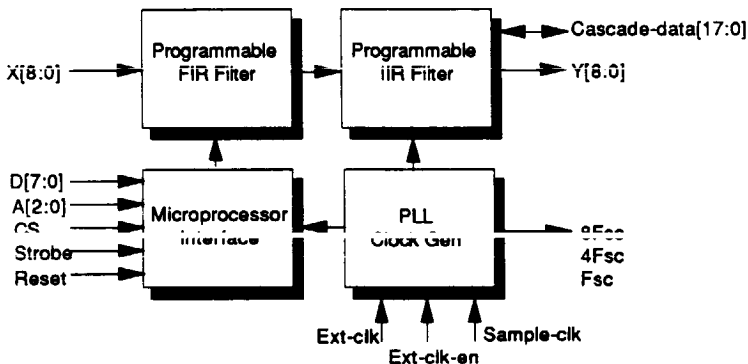
Revision 1.1
16 May 1991

Introduction

The Philips Ghostbuster GB-180 is a 180 tap configurable IIR/FIR filter designed specifically for the cancellation of TV ghosts. It is divided into 9 sections of 20 contiguous taps each of which can be placed anywhere in a TV line period by virtue of 9 programmable delay lines. The chip can operate in isolation or may be cascaded for more than 180 filter sections. The chip contains over 430,000 transistors operating at a frequency of 16X Fsc. An on-chip phase-locked loop allows an input clock of 4X Fsc (14.32 MHz) to generate the internal clock of 56 MHz.

Features

- 180 configurable filter stages
- 9 sections of 20 taps
- 9 programmable delay lines (0-511 IIR) (0-191 FIR)
- 8 bit filter sample resolution; 9 bit resolution (typically)
- 18 bit internal and cascade data
- 4X Fsc clock (internal 16X Fsc PLL)
- 18 bit cascade option
- Microprocessor Interface for Coefficients and Setup
- Optional 3.3V operation for low power

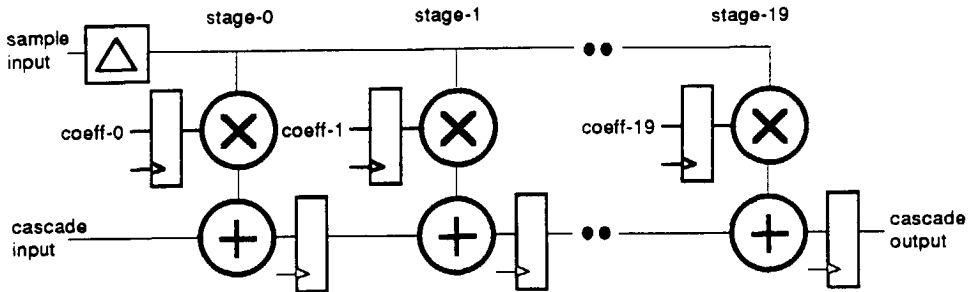


Functional Block Diagram - GB-180 Ghostbuster Chip

Above is shown a block diagram of the chip. It notionally consists of a reconfigurable IIR and FIR filter, a microprocessor interface, output signal conditioning and clock generation. PLL derived clocks are made available for external use.

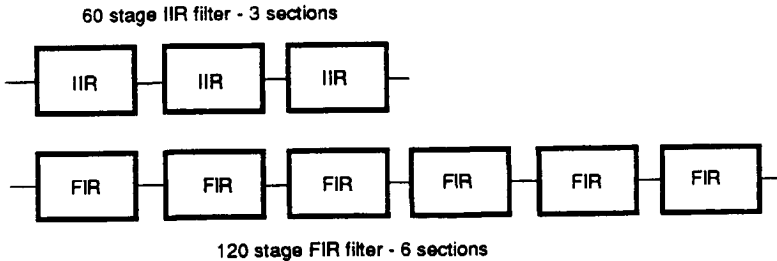
Chip Description

The chip consists of 19 configurable filter *sections*. Each filter section is composed of 20 filter *stages* with loadable coefficients. The figure below shows a section and the internal structure of a stage. Each stage forms a fully programmable filter tap. A variable delay on the front of the filter allows the filter response to be placed anywhere within 512 samples of the start of a line. The sample input is 8 (1.7) bits and the cascade sums carry 18 bits of precision (4.14).



Block Diagram - Filter Section

The sections may be reconfigured to yield a variety of filters. In the figure below, three sections have been combined into an IIR filter, while the remaining 6 sections form an FIR filter. Each particular filter section can be placed anywhere within the active TV line by virtue of programmable delay lines. Configuration is accomplished via the microprocessor port. A final adder allows for rounding and limiting.



Block Diagram - Filter Configuration

Signal Pin definitions

Logical Port	PLCC/PGA Pins
Sample Port:	21/41 Pins

X<8:0> Active High Input

The X input is the sample input for the chip, and is clocked by the rising edge of SAMPLE-CLK. In a single chip implementation, or for the main chip of a multi-chip implementation, this is the input to the filter from the D/A converter. For the cascade chip for a multi-chip implementation, X<8:1> is the filter output Y<8:1> of the main chip, and X<0> is the CASCADE-X<0> of the main chip.

Y<8:0> Active High Output

The Y output is the sample output for the filter, and is clocked by the rising edge of SAMPLE-CLK. In a single chip implementation or for the main chip of a multi-chip implementation this is the output to the D/A converter or to further processing logic. For the cascade chip of a multi-chip implementation, Y is undefined.

CASCADE (PGA only) Active High Input

The CASCADE signal specifies the kind of filter implemented and what the role of this chip is. In a single-chip implementation or for the main chip of a multi-chip implementation, this pin is deasserted. For the cascade chip of a multi-chip implementation, this signal is asserted, and enables the driving of the CASCADE-DATA bus. This signal is level sensitive and is not clocked by SAMPLE-CLK. In the PLCC package, this chip is internally deasserted.

CASCADE-X<0> (PGA only) Active High Output

The CASCADE-X<0> output is used to convey rounding information to cascade chips of multi-chip implementations. This output has the same timing as the Y output. In a multi-chip implementation, this output is used to replace the Y<0> output of the main chip as the input to the X<0> of the cascade chip.

CASCADE-DATA<17:0> (PGA only) Active High Bidirectional

The CASCADE-DATA signals are used to convey information from the cascade chip of a multi-chip implementation to the main chip. CASCADE-DATA is an input to the main chip and is sampled at the rising edge of SAMPLE-CLK. CASCADE-DATA is an output from the cascade chip, and is advanced by rising edge of SAMPLE-CLK.

SAMPLE-CLK Active High Input

The SAMPLE-CLK input clocks all inputs and outputs of the Sample Port. The active edge of this clock is the rising edge. This clock must be extremely stable (need a spec on this) and is used by the internal 4x PLL as its reference clock. The duty cycle of this clock should be unimportant.

EXT-CLK Active High Input

The EXT-CLK input is a clock which is used to bypass the internal 4X PLL for testing purposes and in some system configurations. The active edge of this clock is the rising edge, and the duty cycle must be very close to 50%. There must be a stable phase reference between EXT-CLK and SAMPLE-CLK such that the input setup and hold times are for SAMPLE-CLK are met (these need to be specified).

EXT-CLK-EN Active High Input

The EXT-CLK-EN input is used to select the source of the internal 4X clock. When asserted, the internal clock is taken from the EXT-CLK pin. When deasserted, the internal clock is derived from the SAMPLE-CLK pin by an internal PLL.

Microprocessor Port: 15/15 Pins

D<7:0> Active High Bidirectional

The D signal is used to read and write data from the chip. Currently only the write direction is really defined, and the read direction will probably be used only for testing purposes. The write data is sampled at the rising edge of the STROBE signal.

READ Active High Input

The READ input specifies if the microprocessor cycle is a read or a write. This input is sampled at the rising edge of the STROBE signal.

A<2:0>
CS Active High Input

The A, and CS inputs are used to determine if the current microprocessor cycle is addressing this chip. These inputs are all sampled on the rising edge of the STROBE signal. The chip is selected only if CS is asserted, and A<2> is equal to the CASCADE input. If this chip is selected, then A<1:0> specify which internal register is read or written. The CASCADE input is used to differentiate between the main and cascade chip in multi-chip implementations. The CASCADE input is asserted for one of the chips and deasserted for the other. I

RESET Active High Input

The RESET signal resets the internal state of the chip, and the clock generators. All filter section configuration registers are reset to a known state, and the input clock generator is reset and waits for the first SAMPLE-CLK before starting up. All microprocessor requests in progress are aborted and ignored. There is a minimum pulse width specification needed, and some minimum delay before the next microprocessor cycle can be started.

STROBE Active High Input

The STROBE signal clocks all the signals in the Microprocessor Port except for RESET. The active edge of this clock is the rising edge, and the duty cycle is unimportant. Since there is no handshaking for the microprocessor cycles, there is a maximum repetition rate which will have to be specified (in terms of the 4x clock).

Clock/Status Pins:		0/4 Pins
OVERFLOW (PGA only)	Active High	Output
Set when overflow occurs in final adder. Cleared by CLEAR-STATUS-BIT in global configuration register.		
3.58 MHz (PGA only)	Active High	Output
Nominally 3.58 MHz divided down from internal PLL		
14.32 MHz (PGA only)	Active High	Output
Nominally 14.32 MHz divided down from internal PLL		
28.64 MHz (PGA only)	Active High	Output
Nominally 28.64 MHz divided down from internal PLL		

Power/Ground/Unused Pins:		48/84 Pins
FILTER-VDD/VSS	VDD/VSS	36/36 Pins
FINAL-VDD/VSS	VDD/VSS	2/6 Pins
CLOCK-DRIVER-VDD/VSS	VDD/VSS	2/2 Pins
PLL-VDD/VSS	PLL-VDD/PLL-VSS	2/2 Pins
PAD-VDD/VSS	VDD/VSS	6/14 Pins
Unused (PGA only)	None	0/24 Pins
Total Pins:		84/144 Pins

Microprocessor registers

The microprocessor interface is a write-only interface, with four 8-bit internal registers.

The four registers are addressed by the A<1:0> bits and are:

A<1:0>	Register
0	COEFFICIENT Bits Usage <7:0> Signed coefficient
1	CONFIGURATION-LOW Bits Usage <7:0> CONFIGURATION-DATA<7:0>
1	CONFIGURATION-HIGH Bits Usage <7:4> CONF-SEL<3:0> <3:0> CONFIGURATION-DATA<11:8>
3	TEST (Undefined)

When the COEFFICIENT register is written, and the filter chip is in LOAD-COEFFICIENT mode, then the coefficient shift-register is advanced, and the new data is written into the first element of the shift-register. See the section on loading coefficients for details.

When the CONFIGURATION-LOW register is written, the data written is stored in a holding register which is used later to supply the bottom 8 bits of the CONFIGURATION-DATA being written. This holding register is overwritten when a COEFFICIENT register write is performed, so these operations cannot be interleaved.

When the CONFIGURATION-HIGH register is written, a write to a specific configuration register is triggered. The data for the write comes from the concatenation of the bottom 4 bits of the CONFIGURATION-HIGH write data and from the the CONFIGURATION-LOW holding register. The configuration register to be written is selected by the top 4 bits of the CONFIGURATION-HIGH write data.

CONF-SEL	
0	Filter Section 0 configuration register
1	Filter Section 1 configuration register
2 - 8	Filter Section 2-8 configuration registers
13	Global configuration register 1
14	Global configuration register 0
15	Filter Section 0-8 configuration registers

(All configuration registers are loaded with the same data)

Format of the configuration registers:

Filter Section configuration register:

Bits	Usage
<8:0>	Input Delay (0-511 samples)
9	FIR Enable

Input is from FIR delay line
If output-enable drive FIR-SUM bus

- | | | |
|----|----------------|---|
| 10 | Output enable | Drive FIR-SUM or IIR-SUM bus |
| 11 | Enable Cascade | If zero, the sum input to this section is 0, otherwise it is the sum from the previous filter section |

Global configuration register 0:

- | Bits | Usage |
|-------|--|
| <2:0> | FIR-SCALE (0-4 -> 1/1 to 1/16) |
| <5:3> | IIR-SCALE (0-4 -> 1/1 to 1/16) |
| 6 | BYPASS (bypass output) |
| 7 | 8-Bit Filter (enable 8 bit samples) |
| 8 | Signed samples (assume samples are signed) |
| 9 | LOAD-COEFFICIENT mode |
| 10 | LOAD-COEFFICIENT strobe |
| 11 | Clear Status Bits |

Global configuration register 1:

- | Bits | Usage |
|------|--------------|
| <0> | FIR-SUM zero |
| <1> | IIR-SUM zero |

The TEST register is currently undefined.

Loading coefficients

Naming conventions:

Filter section 0 is the filter section which takes its cascade input from the CASCADE-DATA pins. Filter section 1 is the filter which takes its cascade input from Filter section 0's cascade output, and so on. Filter stage 0 is the first stage in filter section 0, whose sum output is connected to the sum input of filter stage 1. As there are 20 filter stages in each filter section and there are 9 filter sections, there are 180 filter stages in each chip, arranged as follows:

Filter Section	Filter stages
0	0 - 19
1	20 - 39
2	40 - 59
3	60 - 79
4	80 - 99
5	100 - 119
6	120 - 139
7	140 - 159
8	160 - 179

All coefficients are loaded in parallel from a shift register when the LOAD-COEFFICIENT strobe is set in the Global configuration register. The shift register is formed by the filter sum pipeline when the LOAD-COEFFICIENT mode bit is set. The new coefficient is shifted into filter stage 0 when the COEFFICIENT register is written, and all other coefficients are shifted upward. Thus the filter stage coefficients should be written from stage 179 down to stage 0. A schematic program for loading the coefficients would be:

- 1) Write all Filter Section configuration registers to set ENABLE-CASCADE.
- 2) Write the Global Configuration register to turn on LOAD-COEFFICIENT mode.
- 3) Write filter stage 179 coefficient into the COEFFICIENT register
- 4) Write filter stage 178 coefficient into the COEFFICIENT register
- 5) Write filter stage 177->0 coefficients into the COEFFICIENT register
- 6) Write the Global Configuration register to turn on LOAD-COEFFICIENT strobe (leave LOAD-COEFFICIENT mode asserted)
- 7) Write the Global Configuration register to turn off LOAD-COEFFICIENT strobe (leave LOAD-COEFFICIENT mode asserted)
- 8) Write the Global Configuration register to turn off LOAD-COEFFICIENT mode
- 9) Reload all Filter Section configuration registers with the proper values.

Electrical Characteristics

Absolute Maximum Ratings

Ambient Operating Temperature	0-85 C (junction)
Ambient Storage Temperature	-65 - 150 C
Supply Voltage	-0.5 to 7.0 V
Applied Input Voltage	-0.5 to 7.0V
Applied Output Voltage	-0.5 to VDD+0.3V
Power Dissipation (5V)	TBD
(3V)	TBD

Package Thermal Characteristics

TBD

DC Electrical Characteristics

(Temp = 0-85 C)

Parameter	5.0V	3.3V	Units	Comments
LOGIC INPUTS				
V _{ih} , Input High Voltage	TBD		Volts	
V _{il} , Input Low Voltage	TBD		Volts	
C _{in} , Input Capacitance	TBD		pF	
LOGIC OUTPUTS				
V _{oh} , Output High Voltage	TBD		Volts	
V _{ol} , Output Low Voltage	TBD		Volts	
I _{oh} , Output High Current	TBD		µA	
I _{ol} , Output Low Current	TBD		mA	
POWER SUPPLIES				
V _{th}	TBD		Volts	
I _{dd}	TBD		mA	
I _o	TBD		mA	Output Leakage
V _{pll}	5.0V	5.0V	Volts	PLL Analog Voltage

AC Electrical Characteristics

(Temp = 0-85 C VDD ± 5.0V or VDD ± 3.3V)

Output Valid, Output Driven Measurements 50pF load, add .1nS/pF

Parameter	Description	5.0V		3.3V		Units
		min	max	min	max	
						nS
Txs	X<8:0> setup (s)	TBD				nS
Txh	X<8:0> hold (s)	TBD				nS
Tyd	Y<8:0> delay (s)	TBD				nS
Tc0	Cascade-X<0> delay(s)	TBD				nS
Tcd	Cascade-Data<17:0> delay (s)	TBD				nS
Tcs	Cascade-Data<17:0> setup (s)	TBD				nS
Tch	Cascade-Data<17:0> hold (s)	TBD				nS
Tss	Sample-Clk setup (e)	TBD				nS
Tsh	Sample-Clk hold (e)	TBD				nS
Tsd	Sample-Clk delay (e)	TBD				nS
Toe	Ext-Clk period	TBD				nS
Tds	D<7:0> setup (st)	TBD				nS
Tdh	D<7:0> hold (st)	TBD				nS
Tas	Read, A<2:0>,CS setup (st)	TBD				nS
Tah	Read, A<2:0>,CS hold (st)	TBD				nS
Tres	Reset setup (s)	TBD				nS
Treh	Reset hold (s)	TBD				nS
Tcs	Strobe period	TBD				nS

Notes:

(s) referenced to Sample-Clk rising

(e) referenced to Ext-clk rising

(st) referenced to Strobe rising

Packaging

84 pin PLCC

Package Diagram



Pin Assignments

Pin	Signal	Pin	Signal
1	X<8>	17	VSS
2	X<7>	18	SAMPLE CLOCK
3	X<6>	19	VDD
4	X<5>	20	VSS
5	X<4>	21	VDD
6	X<3>	22	VSS
7	X<2>	23	VDD
8	X<1>	24	VSS
9	X<0>	25	VDD
10	VDD	26	VSS
11	VSS	27	VDD
12	PLL-VDD (5V)	28	VSS
13	PLL-VSS	29	VDD
14	VDD	30	VSS
15	VSS	31	VDD
16	VDD	32	VSS

33	VSS	59	VDD
34	VDD	60	VSS
35	D<0>	61	VDD
36	D<1>	62	VSS
37	D<2>	63	VDD
38	D<3>	64	VSS
39	D<4>	65	VDD
40	D<5>	66	VSS
41	D<6>	67	VDD
42	D<7>	68	EXT-CLK
43	Y<0>	69	VSS
44	Y<1>	70	VDD
45	Y<2>	71	VSS
46	Y<3>	72	VDD
47	Y<4>	73	VSS
48	Y<5>	74	VDD
49	Y<6>	75	VDD
50	Y<7>	76	VSS
51	Y<8>	77	STROBE
52	VSS	78	RESET
53	VDD	79	CS
54	VSS	80	A<2>
55	VDD	81	A<1>
56	VSS	82	A<0>
57	VDD	83	READ
58	VSS	84	EXT-CLK-EN

SUMMARY

This thesis starts, after a short introduction, in chapter 2 with the description of a few fundamentals of current conventional television technology and the occurrence of echoes. Chapter 3 gives a preliminary review of the character of these echoes. After that, a more complete discussion is devoted in chapter 4 to describing how echoes are generated and how they affect the received base-band video signal in various modulation systems. An important conclusion is that when viewed from baseband at the transmission end through baseband at the receiving end, the process of echo creation is linear. Based on this insight, a list of requirements for a GCR (Ghost Cancellation Reference signal) is set up in chapter 5. This again leads in chapter 6 to the actual design of a family of GCRs for universal use in the various TV systems in the world. Our considerations up to that point focus entirely on an analog GCR and therefore the potential of a digital GCR has to be investigated separately, which is done in chapter 7, reaching the conclusion that an analog GCR of the kind that we designed is preferable. The use of a GCR for channel characterization is the subject of chapter 8 and the final step, a treatment of actual filter use for echo cancellation is taken in chapter 9, followed in chapter 10 by a system overview in terms of algorithms and an account of our computer simulations on cancellation, which concludes the theoretical part of this thesis.

In the next chapter, number 11, we contemplate more practical matters: the hardware design of a real echo cancellation system. The complementary text on software design constitutes chapter 12. Chapter 13 sums up the various reasons for malfunctioning in the system that may occur.

We report on subsequent testing process in the USA and the pertinent test results in chapter 14. We underwent these in competition with four other systems among whom our system survived eventually as the winner in the USA: it became a national standard. After that chapter 15 describes some aspects of transcending the national scene towards the international scene.

Chapter 16 gives some conclusions on the work reported in this thesis and finally chapter 17 mentions some prospects for future work.

SAMENVATTING (SUMMARY IN DUTCH)

Dit proefschrift begint, na een korte inleiding, in hoofdstuk 2 met de beschrijving van enige fundamentele aspecten van conventionele televisietechnologie en het fenomeen van echo's. Hoofdstuk 3 geeft een voorlopig overzicht van het karakter van deze echo's. Daarna wordt in hoofdstuk 4 een meer complete discussie gewijd aan de creatie van echo's en aan de vraag hoe ze het ontvangen videosignaal aantasten op basisbandniveau onder het gebruik van verschillende modulatiesystemen. Een belangrijke conclusie is dat, wanneer beschouwd van basisband aan de zenzijde tot basisband aan de ontvangzijde, het echo-creatieproces lineair is. Steunend op dit inzicht biedt hoofdstuk 5 een lijst van eisen die aan een echo-onderdrukking-referentie (EOR) signaal moeten worden gesteld. Dit leidt dan in hoofdstuk 6 op directe wijze tot het werkelijk ontwerp van een familie van EOR's voor universeel gebruik in de verschillende televisiesystemen die in de wereld dienst doen. De overwegingen zijn tot aan dit stadium geheel geconcentreerd op een analoge EOR, zodat de mogelijkheden van een digitale EOR apart onderzocht dienen te worden. Dit gebeurt in hoofdstuk 7 waarbij geconcludeerd wordt dat een analoge EOR van de soort die wij ontwierpen te prefereren is. Het gebruik van een EOR voor kanaalkarakterisering is het onderwerp van hoofdstuk 8 en de uiteindelijke stap naar het werkelijk programmeren van digitale filters voor echo-onderdrukking wordt genomen in hoofdstuk 9, gevolgd in hoofdstuk 10 door een systeemoverzicht vanuit algoritmisch standpunt en een verslag over onze computersimulaties van echo-onderdrukking, hetgeen het besluit vormt van het theoretisch gedeelte van dit proefschrift.

In het volgende hoofdstuk, nummer 11, beschouwen we een meer praktisch

onderwerp, het "hardware"-gedeelte van het door ons ontworpen onderdrukkingssysteem. De corresponderende tekst over de "software" is het onderwerp van hoofdstuk 12. Hoofdstuk 13 is een opsomming van de verschillende redenen waardoor het systeem soms minder goed functioneert.

Hoofdstuk 14 rapporteert over de ondergane testen in de VS en de bijbehorende testresultaten. Ons systeem onderging deze tests in concurrentie met 4 andere systemen en overleefde tenslotte als winnaar in de VS. Het werd een nationale standaard. Hierna beschrijft hoofdstuk 15 enkele aspecten van de transitie van de nationale naar de internationale scène, die eindigde met internationale standaardisering van ons systeem met uitzondering van Japan en Korea.

Hoofdstuk 16 biedt enkele conclusies over het werk dat beschreven wordt in dit proefschrift en tenslotte noemt hoofdstuk 17 enkele mogelijkheden voor toekomstig werk.

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BIOGRAPHY

David Koo was born in Shanghai, China. He graduated from the No.04 Department (Electrical Engineering Department) of the Shanghai University of Science and Technology, Shanghai, China. He worked in the Shanghai Television Technology Research Institute as a Researcher and an Engineer for more than four years. He later graduated with a Masters degree from the Electrical Engineering Department of C.C.N.Y of the City University of New York, NYC, NY, U.S.A.

David Koo joined Philips in 1984, and since then has been working at Philips Laboratories at Briarcliff, New York, U.S.A., serving as an Associate Researcher, Researcher, Senior Researcher, Principal Researcher and Consulting Scientist. His main field is Signal Processing. He has been working on Echo Cancellation research work since late 1987 and received an Emmy Award for the Echo Cancellation work for television application in 1995.

