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ESTIMATION OF THE IMPULSE RESPONSE OF TELEVISION CHANNELS FROM DIGITIZED WAVEFORMS

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ESTIMATION OF THE IMPULSE RESPONSE OF TELEVISION CHANNELS FROM DIGITIZED WAVEFORMS

by

A. Vincent, M. Bruneau

ABSTRACT

This report presents a technique to characterize television channels without disrupting the normal video program. TELETEXT data lines transmitted during the vertical blanking interval are digitized. The least-squares analysis is applied on the sampled data to estimate the channel impulse response, from which the amplitude and phase characteristics are computed.

The report begins by a short review of the television and teletext signals. Then, the theory and implementation of least-squares analysis techniques are presented. Also, the signal processing leading to TV channel transfer function evaluation is described. Finally, experimental results are presented.

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ESTIMATION OF THE IMPULSE RESPONSE OF TELEVISION CHANNELS FROM DIGITIZED WAVEFORMS

1. Introduction

Teletext is a digital communication system, whereby alphanumeric and graphic information is encoded in a digital format, and transmitted during unused lines of a television signal. Television channels introduce impairments such as amplitude and phase distortions, as well as echoes, all of which produce intersymbol interference and decrease the noise margin.

In order to characterize data transmission over TV channels, it is important to have a good knowledge of the characteristics of the transmitting and receiving equipment, as well as of the propagation medium. To do this, the standard television test signals (1T, 12.5T, multiburst, etc.) are often used. However, they provide insufficient information to fully characterize the channel, and they do not always originate from the same source as the teletext signal. The amplitude and group delay could be measured with specialized test equipment, such as a network analyzer. This requires that measurements be made during off-hours, in order to have the entire TV channel available, which is not too practical, especially if a large number of measurements are to be made.

This report describes a system which allows estimation of the impulse response of TV channels, based on least-squares analysis (LSA), using digitized teletext waveforms. The transfer function of the channel, through which the teletext signal is transmitted, can therefore be measured without disrupting the normal video program and without using special equipments at the transmit end.

In section 2, the television and teletext signals are briefly reviewed. Section 3 introduces the theory of impulse response estimation, based on least-squares analysis. In section 4, the implementation of the least-squares techniques for the estimation of the impulse response of TV channels is described, as well as the transmitting and receiving systems.

A description of the signal processing leading to the system transfer function is described in section 5, whereas section 6 presents results. The conclusion is given in section 7.

SECTION 2. TELEVISION AND TELETEXT SIGNALS

2.1 Television Channel

A block diagram of a broadcast television system is shown in figure 1 (the audio signal is not shown). Vestigal Sideband (VSB) modulation is used to conserve spectrum; i.e., the lower sideband is only partially removed. The transmitted signal is double sideband (DSB) over the region of 0-0.75 MHz from the visual carrier, and single sideband (SSB) in the region 1.25 MHz. and above. The band 0.75 to 1.25 MHz is a transition band from DSB to SSB. The transmitted signal is therefore not vestigial sideband; it is filtering at the receiver, which creates the VSB signal (fig. 1(d)). The audio sub-carrier is at 4.5 MHz above the visual carrier. The video signal is limited to approximately 4.2 MHz in order to avoid interference with the audio signal.

Broadcast television channels may be represented by an equivalent baseband complex impulse response (see Appendix A):

$$h(t) = h_I(t) + jh_Q(t) \qquad (2.1)$$

where $h_I(t)$ and $h_Q(t)$ are respectively the in-phase (I) and quadrature (Q) baseband components of the impulse response. Their Fourier transforms:

$$\begin{aligned} h_I(t) &\leftrightarrow H_I(f) \\ h_Q(t) &\leftrightarrow H_Q(f) \end{aligned} \qquad (2.2)$$

represent the transfer function of the baseband in-phase and quadrature channels, depicted in Figures 2 (a) and (b), respectively.

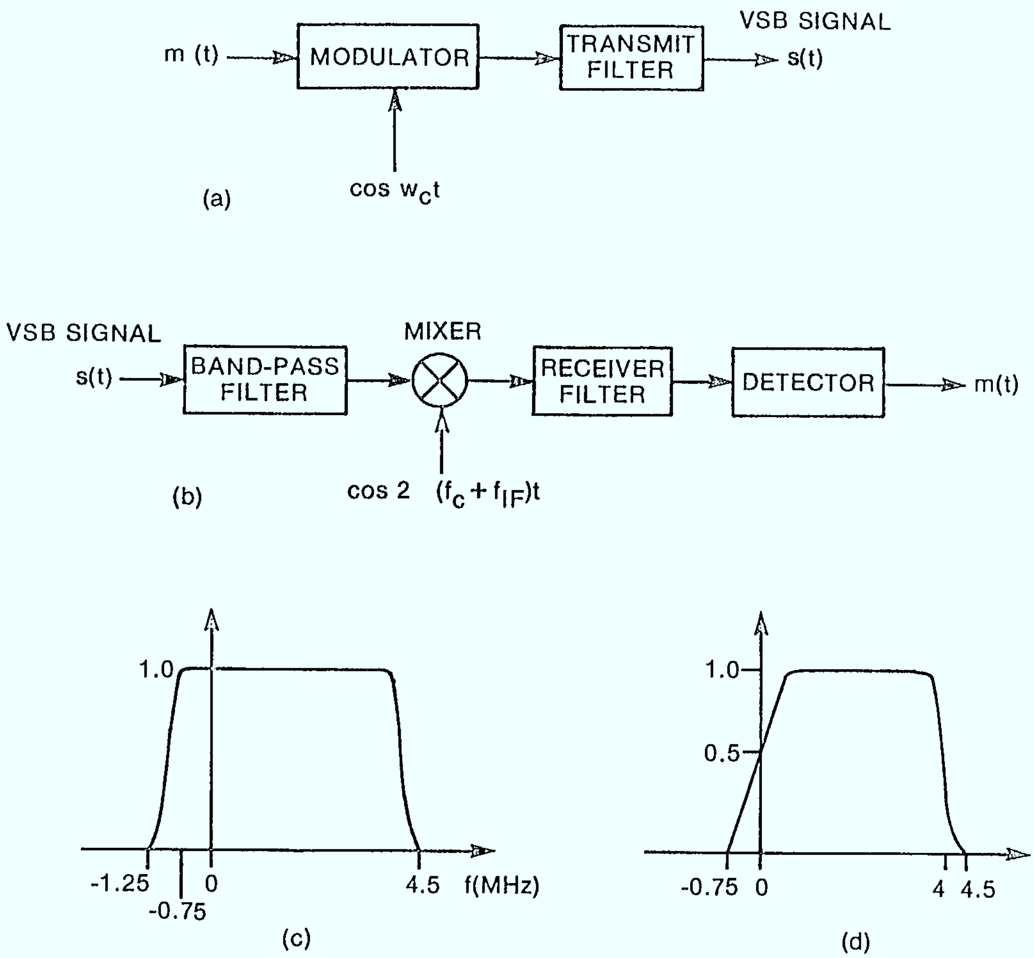


Figure 1: Simplified Block Diagram of Television Broadcast System:
(a) VSB modulator
(b) VSB demodulator
(c) magnitude response of transmit filter
(d) magnitude response of receiving filter.

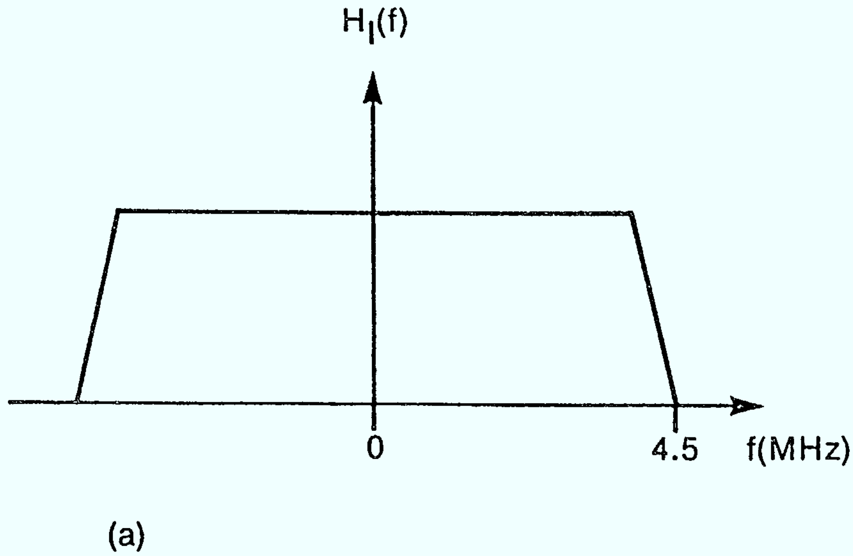


Figure 2 (a): Equivalent baseband in-phase TV channel

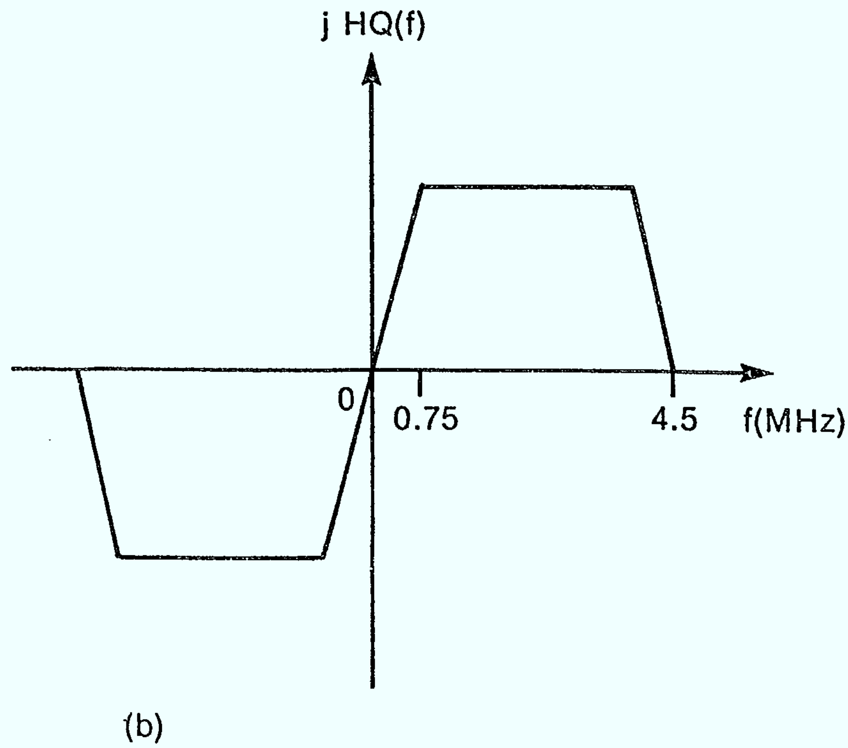


Figure 2 (b): Equivalent baseband in-quadrature TV channel

Television channels can be characterized by their complex baseband impulse response. This complex response may be obtained from the in-phase and the quadrature outputs of a synchronous quadrature demodulator.

Most present day television demodulators are either of the quasi-synchronous or envelope detector types. Quasi-synchronous detectors employ a carrier recovery bandwidth which is wider than that of synchronous detectors. The recovered carrier may therefore exibit relatively large phase jitters, which would produce quadrature distortion. Envelope detectors, react entirely to both the in-phase and the quadrature signals.

Thus, the output of modern television receivers can only be accurately modelled and analyzed when the complex baseband impulse response is known [13].

The impulse response estimation scheme presented in this report, processes the in-phase and the quadrature components separately, to produce an in-phase and quadrature baseband impulse response. The latter can be combined to yield the complex impulse response.

Due to the unavailability of a quadrature demodulator, during the system development phase, results are shown for the real impulse response only. However, the concept is also applicable to the imaginary part.

2.2 Teletext Signal

The teletext signal format is described in [11]. The data is transmitted at a bit rate of 5.727272 MB/s, using non-return to zero (NRZ) coding, and pulse amplitude modulation (PAM). The data amplitudes are 0 and 70 IRE for levels 0 and 1 respectively. Pulse shaping is raised cosine spectrum, with roll-off between 55% and 100%.

The data packets consist of:

- 2 bytes of alternating 1 and 0 for bit-synchronization
- a byte-synchronization byte
- a 28-byte data field which may include up to 3 bytes for error correction.

The packets are inserted during unused lines of the vertical blanking interval (VBI) of a TV signal. Figure 3 (a) shows a portion of a video signal, with teletext on line 18, and figure 3 (b) shows an expanded view of line 18, showing a typical teletext packet.

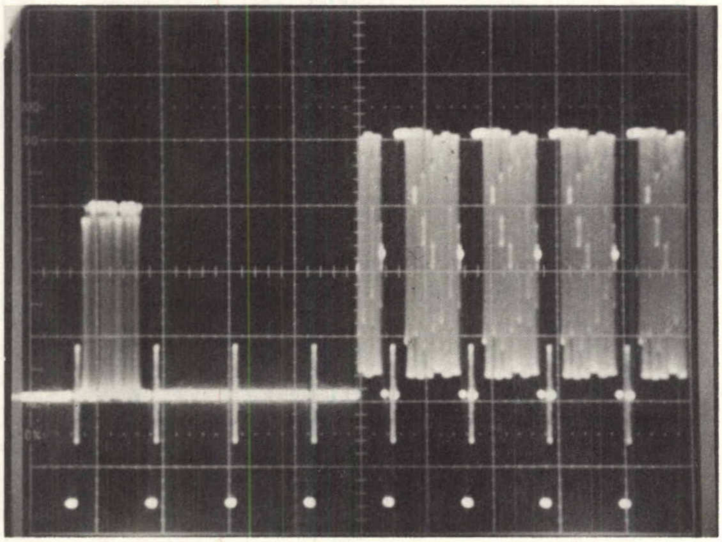
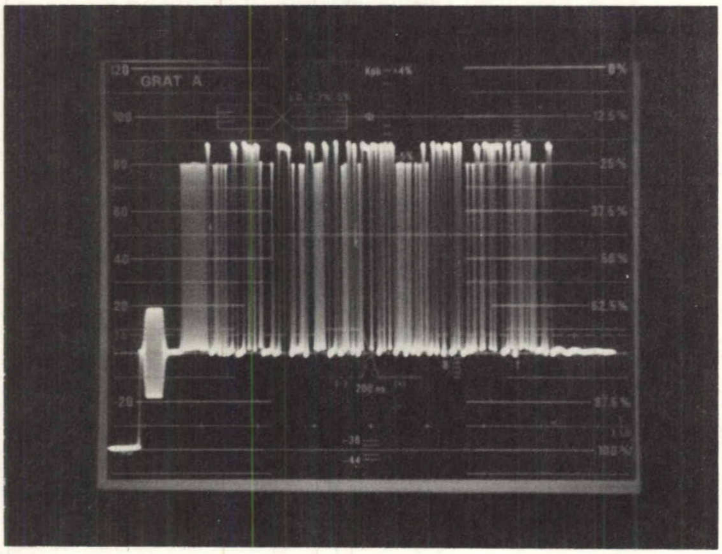


Figure 3 (a): Portion of a television signal



(b)

Figure 3 (b): Portion of a teletext line

SECTION 3. THEORY OF IMPULSE RESPONSE ESTIMATION

3.1 Impulse Response

The television channel can be assumed to be linear and time-invariant. It can, therefore, be modeled as the equivalent baseband model shown in figure 3.1. The input x(t) and output y(t) are then related by the following expression:

$$y(t) = x(t) * h(t) + n(t) \tag{3.1}$$

where * denotes the convolution integral. Note that h(t) and y(t) are in general complex. The equivalent baseband complex impulse response h(t) represents the response of the overall channel, consisting of the transmit and receive filters, as well as the propagation channel.

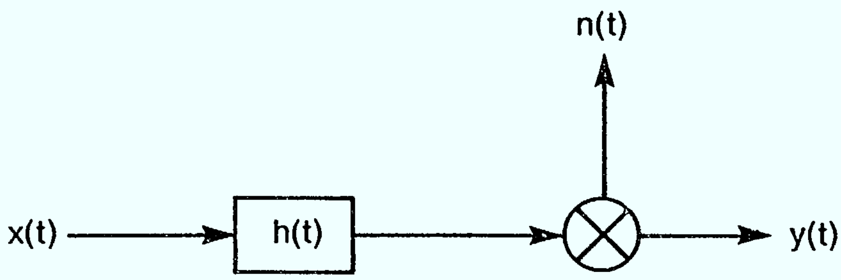


Figure 3.1: Baseband model.

The simplest method of computing the impulse response is by spectral division. If we take the Fourier transform on both sides of Eq. (3.1):

$$Y(f) = X(f) \cdot H(f) + N(f)$$

then

$$H(f) = \frac{Y(f)}{X(f)} - \frac{N(f)}{X(f)}$$

Aside from the practical issue of windowing, the above estimate is slow to converge, and is subject to large errors (noise enhancement) when X(f) is small (e.g., when the input signal is band-limited [6]. Methods such as Modified Periodigrams [5], Short-Time Spectral Analysis [6] and other spectral analysis techniques have been derived to provide more robust estimates in the presence of noise.

3.2 Least-Squares Estimation

The problem of determining the impulse response of a system, may be resolved with the estimation theory to produce an optimum estimate of the impulse response. One of the most widely used estimators in system identification is the Least-Squares (LS) Estimate [3, 4, 7], which under certain conditions corresponds to the Maximum Likelihood Estimate (MLE). In this section, the theory leading to the least-squares estimate is reviewed.

It has been assumed that the television channel is linear time-invariant. Furthermore, the system impulse response h(t) is of finite duration and is causal:

$$h(t) = 0 \text{ for } t<0 \text{ and for } t>\tau;\ \tau>0. \qquad (3.2)$$

Refering to Figure 3.1, the input and output signals can be sampled at a rate 1/T samples/second, equal to the Nyquist rate, or higher.

Due to the linearity assumptions and Eq. (3.2) we can model the channel as a finite impulse response (FIR) filter. The input/output relationship is assumed to be given by the linear regression model:

$$y(i) = \sum_{k=0}^{M} h(k)x(i-k)+n(i) \quad i=0,1,\ldots N \tag{3.3}$$

where x(i) and y(i) represent the set of observations of the input and output sampled signals respectively; h(k) represents the sampled impulse response of duration of M samples; and h(k)=0, for k=0 and $k \geq M$; n(i) represents the measurement noise, and N is the number of samples of x, y and n.

Equation (3.4) represents a set of N linear algebraic equations which may be written in a maxtrix form:

$$\begin{bmatrix} y_o \\ y_1 \\ \cdot \\ \cdot \\ \cdot \\ y_N \end{bmatrix} = \begin{bmatrix} x_o(o) & \ldots & x_o(M) \\ x_1(o) & \ldots & x_1(M) \\ \cdot & & \cdot \\ \cdot & & \cdot \\ \cdot & & \cdot \\ x_N(o) & \ldots & x_N(M) \end{bmatrix} \cdot \begin{bmatrix} h_o \\ h_1 \\ \cdot \\ \cdot \\ h_M \end{bmatrix} + \begin{bmatrix} n_o \\ n_1 \\ \cdot \\ \cdot \\ \cdot \\ n_N \end{bmatrix}$$

or $$Y = XH + N \tag{3.4}$$

The objective is to determine a coefficient vector $B=\{b_o, b_1 \ldots b_M\}$ which estimates H, the system impulse response. This is done by using the theoretical response (W) of a model of the system. Since we have assumed the channel to be linear, we can model it as a transversal filter, as shown in Figure 3.2. The input/output relation of this model is then:

$$w(i) = \sum_{k=0}^{M} b(k)x(i-k) \qquad (3.5)$$

or, in vector-matrix form:

$$W = XB \qquad (3.6)$$

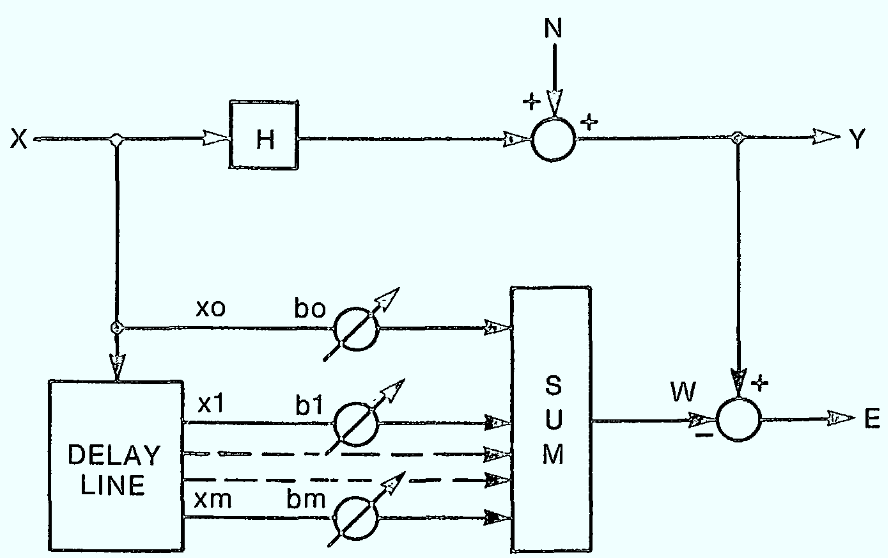


Figure 3.2: System Identification Model

The coefficients $(b_o, b_1, \ldots b_M)$ are then chosen to minimize some function of the measurement error (residuals):

$$e = Y-XB \tag{3.7}$$

$$e' = [e(1), e(2), \ldots e(N)]$$

where the prime denotes the transpose of a matrix.

We digress at this point to discuss about some characteristics of the noise in Equation (3.3). The noise may be given by its mean and covariance matrices:

$$[n'] = [n(1), n(2) \ldots n(N)]$$

$$R = [n.n'] = \begin{bmatrix} [n(1)n(1)] & \ldots & [n(N)n(1)] \\ \cdot & & \\ \cdot & & \\ \cdot & & \\ [n(1)n(N)] & \ldots & [n(N)n(N)] \end{bmatrix}$$

If the noise is Gaussian, no information is lost, since Gaussian density functions are completely characterized by their first and second moments. Note, however, that Gaussian probability density is not a necessary prerequisite for the evaluation of the least-squares procedure, but it actually helps to explain some basic relations.

We shall make the following assumptions:

a) the measurement errors $n(k)$ are independant of each other, and they are randomly distributed, with a normal density $f(n_k)$;

b) the error n(k) and the input signal x(k) are uncorrelated.

These properties can be expressed by:

$$[n(k)\cdot x(k-\tau)] = 0$$
$$[n(k)\cdot n(k-\tau)] = 0; \ \tau\neq 0$$

$$f_n(n_k) = \frac{1}{\sigma_k\sqrt{2\pi}} \quad \exp[-\frac{1}{2\sigma_k^2}\cdot(n_k-\varepsilon[n_k])^2]$$

where σ_k is the standard deviation, and $\varepsilon[\,]$ denotes the expectation of a variable.

As we shall see later, unbiasedness of the estimates requires that

$$\varepsilon[n_k] = 0$$

where the bias is defined as:

$$\text{bias} = \varepsilon[B] - H$$

Due to the above assumptions, we can express the joint probability density function of all measurement errors n_k as [3, 7]:

$$f(n) = \frac{1}{(2\pi)^{k/2}|R|^{\frac{1}{2}}} \quad \exp[-\tfrac{1}{2}n'R^{-1}n]$$

where $R = \varepsilon[nn']$, the covariance matrix of the noise is:

$$R = \begin{bmatrix} \sigma_1^2 & 0 & \dots & 0 \\ 0 & \sigma_2^2 & \dots & 0 \\ 0 & 0 & & \sigma_N^2 \end{bmatrix}$$

Not having any prior knowledge of the noise statistics, we assume that $R = \sigma^2 I$,
where I is the identity matrix, and σ^2 is the noise variance.

This implies that the measurements performed at times k = 1, 2 ... N are corrupted by noise having the same statistical properties, i.e., that all measurements are made with the same precision.

Returning to the least-squares problem, the least-squares method determines the most probable value of B, which minimizes the sum of the square of the errors [3,4,7]:

$$E = [Y-XB]'R^{-1}[Y-XB] = e'R^{-1}e = ||e^*||^2 \tag{3.8}$$

for $R = \sigma^2 I$, we have:

$$\frac{dE}{dB} = -X'[Y-XB] = 0 \tag{3.9}$$

and the least-squares estimates are given by (the so-called normal equation) [3,4,7]:

$$B = [X'X]^{-1}X'Y \tag{3.10}$$

It can be shown [3] that for Gaussian noise, the LS estimates coincide with the Maximum Likelihood estimates having the minimum variance of all unbiased estimates.

The bias of the estimate is given by:

$$\varepsilon[B] = \varepsilon[[X'X]^{-1}X'Y]$$

From Equation (3.4),

$$\varepsilon[B] = \varepsilon[[X'X]^{-1}X'(XH+n)]$$
$$= \varepsilon[[X'X]^{-1}X'XH + [X'X]^{-1}X'n]$$
$$= H + \varepsilon[[X'X]^{-1}X'n].$$

As the input and noise signals are statistically independent:

$$\varepsilon[B] = H + \varepsilon[[X'X]^{-1}X'].\varepsilon[n]$$

Consequently, if $\varepsilon[n] = 0$, the estimate is unbiased, and

$$\varepsilon[B] = H.$$

Another characteristic of the estimate B is its covariance

$$\text{cov}[B] = \varepsilon[(B-H)(B-H)']$$
$$= [X'X]^{-1}X'RX[X'X]^{-1}$$

If the noise is white, $R = \sigma^2 I$ and

$$\text{cov}[B] = \sigma^2[X'X]^{-1} \qquad (3.11)$$

Equation (3.10) may be written in terms of x:

$$\begin{bmatrix} b_o \\ \cdot \\ \cdot \\ \cdot \\ b_M \end{bmatrix} = \begin{bmatrix} \sum_{i=1}^{N} x_o^2(i) & \cdots & \sum_{i=1}^{N} x_o(i)x_M(i) \\ \cdot & & \cdot \\ \cdot & & \cdot \\ \cdot & & \cdot \\ \sum_{i=1}^{N} x_M(i)x_o(i) & \cdots & \sum_{i=1}^{N} x_M^2(i) \end{bmatrix}^{-1} \cdot \begin{bmatrix} \sum_{i=1}^{N} x_o(i)y(i) \\ \cdot \\ \cdot \\ \cdot \\ \sum_{i=1}^{N} x_M(i)y(i) \end{bmatrix}$$

We can define the empirical correlation functions:

$$\tilde{\psi}_{xx(r)} = \frac{1}{N+1} \sum_{n=0}^{N} x(n)x(n+r)$$

$$\tilde{\psi}_{xy(r)} = \frac{1}{N+1} \sum_{n=0}^{N} x(n)y(n+r)$$

These quantities are not correlation functions in the usual sense, but they are calculated directly from the observed finite sequences x(n) and y(n). Therefore,

$$\varepsilon[\tilde{\psi}_{xx(r)}] = \frac{N+1-|r|}{N+1} \cdot \psi_{xx(r)} \qquad (3.12)$$

where $\psi xx(r)$ is the autocorrelation function of x(n). Therefore:

$$\tilde{\psi}_{xx(r)} \to \psi_{xx(r)} \qquad \text{as } N \to \infty$$

Equation (3.10) then becomes:

$$\sum_{k=0}^{M} b(k)\tilde{\psi}_{xx}(k-i) = \tilde{\psi}_{xy}(i) \qquad (3.13)$$

This represents the sampled-data analog of the Weiner-Holpf equation [3,4]. This is the set of equation obtained with the method of "deconvolution".

3.3 Least-Squares Algorithm

The solution of the least squares problem for system identification (Eq. 3.10) requires matrix inversions and multiplications, which are computationally unatractive if the order of the matrices is large. Also, it is not known beforehand how many parameters are significant, (i.e., the length of the impulse response is unknown). It is therefore of interest to have an estimation procedure that is recursive in the number of parameters. This means that the estimate of the parameters $b'(m) = [b_o, b_1, \ldots b_m]$ may be obtained from the estimation of $b'(m-1) = [b_o, b_1 \ldots b_{m-1}]$ without doing all of the calculations again.

Such methods and related algorithms have been extensively reported in the literature. The algorithm used here is due to Marple [9]. This algorithm exploits certain structures of the matrix normal equation, to yield a computationaly efficient algorithm. It starts at order m=0, and recursively computes all order solutions up to m=M, the maximum length of the impulse response.

It should be remembered that least square estimation require that N, the number of observations, be larger than 2M ie:

$$N \geq 2M+1.$$

Further details on the algorithm are given in [9].

3.4 Optimum Inputs

It can be seen from Equations (3.11) that the variance of the estimates becomes smaller as the amplitude of the input sequence x(n) increases. In practical situations, x(n) is limited. Given the constraint that the mean square value of x(n) remains fixed,

$$\tilde{\psi}_{xx(o)} = \frac{1}{N+1} \sum_{n=0}^{N} x^2_{(n)} \tag{3.14}$$

What is the optimum x(n) which minimizes the variance? Assuming white noise, the covariance is given by:

$$\mathrm{cov}[B] = \sigma^2[XX\ ]^{-1}$$

It can be shown [4] that the variance of all b(k) will be minimized if and only if:

$$XX' = (N+1)\tilde{\psi}xx(0)I \qquad (3.15)$$

or

$$\tilde{\psi}_{xx}(0) \neq 0$$
$$\tilde{\psi}_{xx}(r) = 0 \qquad 0<r\leq M.$$

This implies that the sequence x(n) must be white over a range of M sampling intervals. For practical purposes, it is advantageous to use deterministic signals that have the desired autocorrelation function. One such signal, is the maximum length sequence generated by feedback shift registers.

An n tap shift register operating at a clock of period T, generates a sequence of $N = (2^n-1)$ bits, which is periodic, with a period of $(2^n-1)T$.

The autocorrelation function of a maximum length sequence with amplitudes of +a and -a, is shown in figure 3.3.

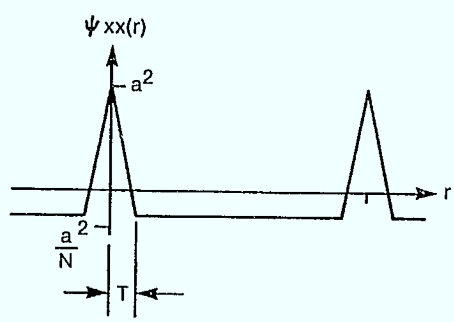


Figure 3.3: Autorocrelation Function of PRBS.

3.5 Noise Error

The received signal, $y(n)$, may contain additive noise, which will inevitably reduce the estimation accuracy.

Let $h(m)$ and $\hat{h}(m)$ be the real and estimated impulse responses, and the coefficient error,

$$\Delta h(m) = h(m) - \hat{h}(m) \qquad m = 0, 1, \ldots M-1$$

The norm of the coefficient error is given by:

$$E = \sum_{m=0}^{M-1} \Delta h^2(m)$$

It can be shown, [6], that for a white input signal $x(n)$ and uncorrelated noise:

$$E \simeq \frac{M}{N} \frac{\sigma_n^2}{\sigma_x^2} \qquad (3.16)$$

where N: number of samples of $x(n)$ and $y(n)$

M: length of impulse response

σ_n^2: variance of the noise

σ_n^2: variance of signal $x(n) = \psi_{xx}(o)$.

Thus, the error is a function of the ratio M/N, and of the input signal-to-noise power ratio.

For each parameter h_j, the standard deviation is given by:

$$\sigma_{hj} = \frac{\sigma_n}{\sqrt{N}\,\sigma_x} \qquad (3.17)$$

which is independent of the number of parameters to be estimated.

The above expression represent a lower bound because, in general, the input signal and noise may not be white and uncorrelated. However, it will be true, in general that $\sigma_{hj} \rightarrow 0$ when N is large, and/or the noise variance tends to zero.

SECTION 4. APPLICATIONS TO TV CHANNELS MEASUREMENTS

4.1 Modelling

We have seen that the LS method provides unbiased estimation, provided certain conditions, are satisfied and that the variance of the estimates may be made small when the number of observations of x(n) and y(n) is made large. There are evident advantages to using a teletext signal to make in-band measurements of the impulse response without disrupting the normal video program. There are, however, severe constraints associatied with this scheme:

a) teletext packets are of rather short duration: 240 bits in the present system, of which the first 3 bytes (bit and byte synchronization) are deterministic;

b) the duty cycle is low: 60 lines/second;

c) the transmitter and receiver are not co-located, so that the input and output signals cannot be sampled simultaneously, and there is not an absolute time reference between them;

d) the teletext signal has a DC bias;

e) the signal is band-limited and is not white within the pass-band.

The packet length prevents the use of long observation intervals to reduce the variance of the estimates. Fortunately, the signal-to-noise ratios encountered in practice are generally high (typically better than 30 dB of peak video to rms noise ratio). Because the test signal is not white, the estimates are less accurate at the high end of the spectrum, where there is less signal energy [6].

The input sequence x(n) may be locally regenerated from y(n), since the transmitted sequence is known. This may be done by simply slicing y(n). This produces a sampled square wave representing the input signal. The pulse shaping filter is considered to be part of the overall channel. However, this square wave, not being band-limited, cannot be sampled at the Nyquist rate without aliasing. The input sequence is therefore modelled as an impulse train:

$$x(n) = \sum_{n=o}^{N} a_n \delta(t+nT)$$

where the $\{a_n\}$ are estimated from the amplitude values of y(n). Since the transmitted sequence is known, possible errors in the estimated $\{a_n\}$ may be corrected.

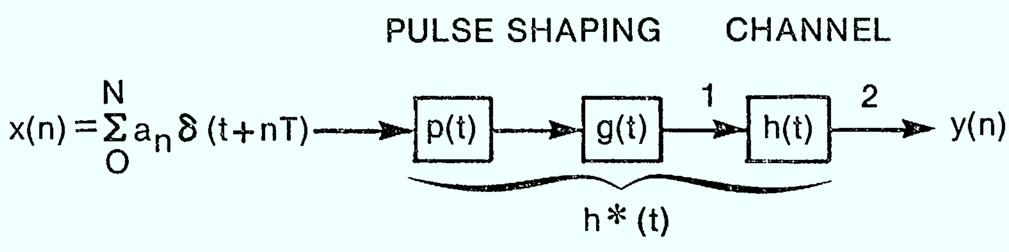


Figure 4.1: Total Channel Model.

The total channel model is shown in Figure 4.1 where:

- $p(t)$ shapes the input signal $x(n)$, which is modelled as an impulse train, into of a NRZ pulse train; thus

$$p(t) = 1 \qquad |t| \leq T/2$$
$$\quad = 0 \qquad |t| > T/2$$

and $$P(w) = F[p(t)] = \frac{\text{Sin } (W\ T/2)}{W\ T/2}$$

- $g(t)$ is the impulse response of the pulse shaping filter i.e., the Raised Cosine filter currently being used in teletext.

- $h(t)$ is the channel impulse response.

The total channel impulse response, $h^*(t)$, is therefore given by

$$h^*(t) = F^{-1}[P(w).G(w).H(w)] \qquad (4.1)$$

The channel impulse response $h(t)$ may be obtained from $h^*(t)$ by spectral division:

$$h(t) = \frac{F^{-1}[P(w).g(w).H(w)]}{F^{-1}[P(w).G(w)]} \qquad (4.2)$$

In other words, the channel impulse response $h(t)$ is estimated in two steps:

a) estimate the impulse response at the output of the pulse shaping filters i.e., at (1).
b) estimate the impulse response of the total channel at (2).
c) calculate $h(t)$ from Eq. (4.2).

The impulse response of the pulse shaping filters at (1) needs to be measured once only, since it is fixed during all the channel measurements.

The above model requires a sampling rate which is a multiple of the bit rate. A sampling rate of twice the bit rate or 11.4545 MHz, is chosen. Recalling that TV channels are band-limited to approximatively 4.2 MHz, this sampling rate is effectively above the Nyquist rate. A phase equalized 5 MHz low-pass filter is used at the receive end as a safety measure, and to remove excess noise.

To improve the accuracy of the estimates in noisy signals would require an increase in the length of the observation interval, which is not possible due to the signal format, which limits the length of teletext packets to approximatly 50 μsecs. Improvements may, however, be obtained by averaging several digitized waveform, since the signal is deterministic. Each output signal sample, y(i) is averaged over M samples where M is the number of successive digitizations of y(t):

$$Y_i = \frac{1}{M} \sum_{m=1}^{M} y_m \qquad i=1,\ 2\ \ldots\ N$$

This scheme is valid only if the sampling clock is phase-synchronized with the sampled signal.

Under such conditions the standard deviation of the noise n decreases as the square root of M:

$$\overline{\sigma_m} = \frac{\sigma_n}{\sqrt{M}}$$

This shows a rather slow convergence. However, with noise levels encountered in practice, there is no need to go beyond M=10, which provides an improvement of 10dB, and consequently will improve the standard deviation of the estimates by the same amount (with white noise) (see Eq. 3.17).

Only the variations of the output signal with respect to variations of the input signal must be used for the estimation algorithm. Therefore, the DC values of the signals must be removed. These DC values are estimated by averaging, and substracted from the signal.

$$\overline{y} = \frac{1}{N} \sum_{i=1}^{N} y_i$$

$$y'_i = y_i - \overline{y}$$

Simulations have shown that the estimation is not sensitive to relatively small DC offsets.

4.2 Transmitting System

An ordinary teletext generator is used to transmit the signals used for impulse response measurements. A block diagram of a CRC-developed teletext generator is shown in Figure 4.2. It consists of a 6809-based microcomputer which generates the appropriate data, and transfers it to an encoder. The encoder appends a prefix consisting of two bytes of bit-synchronization (C) and a byte-synchronization byte(F), and sends the data at 5.727272 MB/s, on the selected VBI lines. The data is then passed through a raised cosine shaping filter. A Tektronix 149 video inserter (or similar device) inserts the data onto selected lines of the video program.

4.3 Receiving System

The system used to measure the real impulse response is shown in Figure 4.3.

The composite video signal (demodulator output) is fed through a waveform monitor, which provides a Line Strobe output which is used to trigger the analog to digital (A/D) converter. The video signal is passed through an anti-alaising filter, which is a phase equalized 5 MHz low-pass filter. It is digitized with an 8-bit A/D converter, operating at twice the bit rate, 11.4545 MHz. The sampling clock is derived from a Norpak MK-4 teletext decoder. It is, in effect, 8/5 th of the colour sub-carrier signal, which is regenerated by the decoder, from the video signal. It is then externally (from the decoder) multiplied by 2, to provide the proper sampling rate. This technique ensures that the sampling clock remains phase-synchronized packet to packet, which makes the averaging of successive packets possible.

Once triggered the A/D converter takes 2048 samples and stores them in its internal memory. These samples may then be transfered to an LSI-11 micro-computer, through a special interface, and stored on floppy disks, for later analysis. In fact, only the first 600 samples are transfered. The rest, being samples of the video signal, are discarded.

The test signal used for the measurement of the impulse response consists of the prefix (C,C,F) and 216 bits of a truncated maximum length pseudo-ramdom sequence. Figure 4.4 shows the amplitude spectrum of the pseudo-random sequence, after the 5 MHz LPF, and Figure 4.5 shows the amplitude response of the 5 MHz LPF.

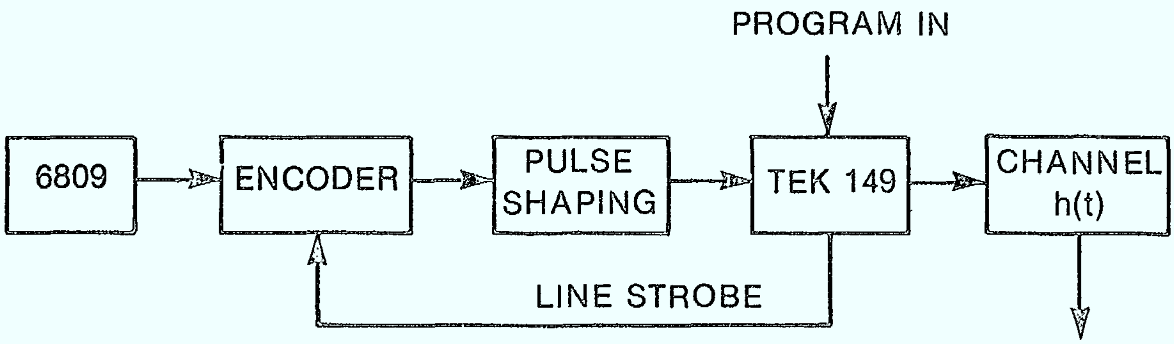


Figure 4.2: Transmitting System.

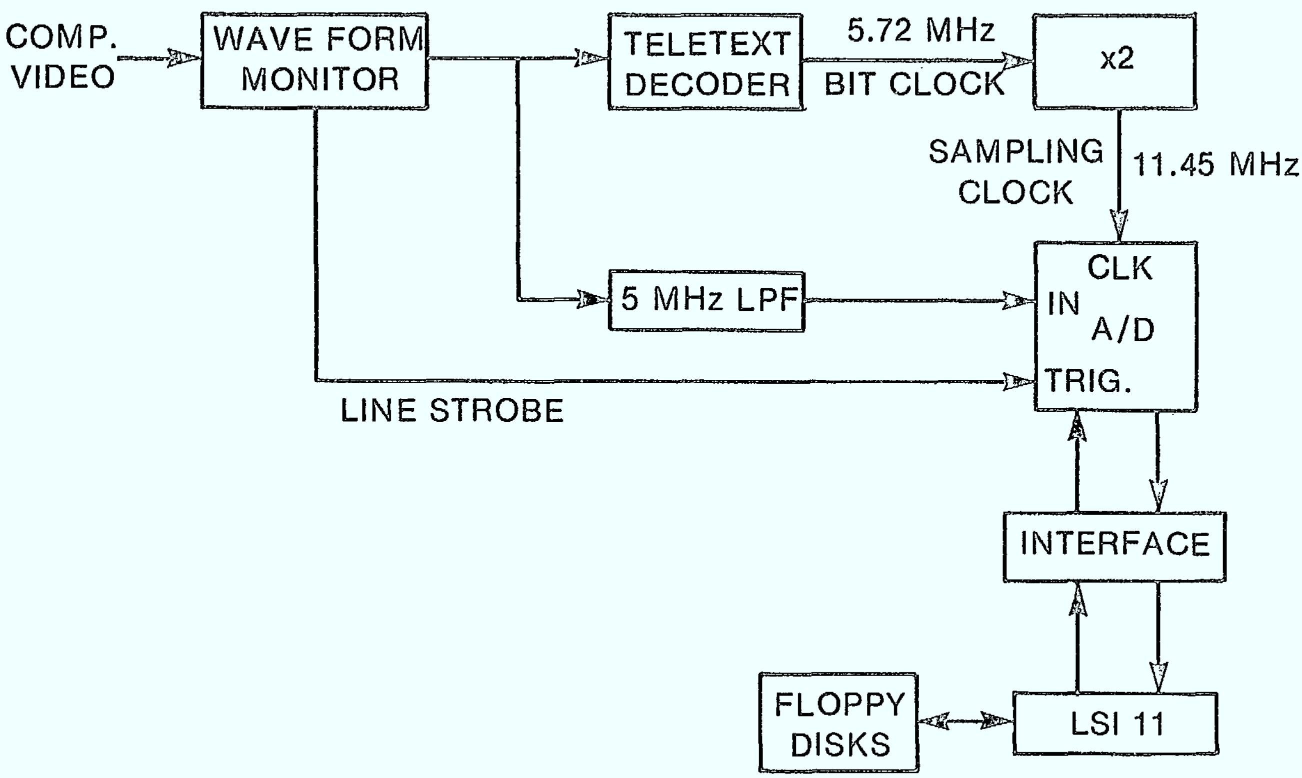


Figure 4.3: Receiving System.

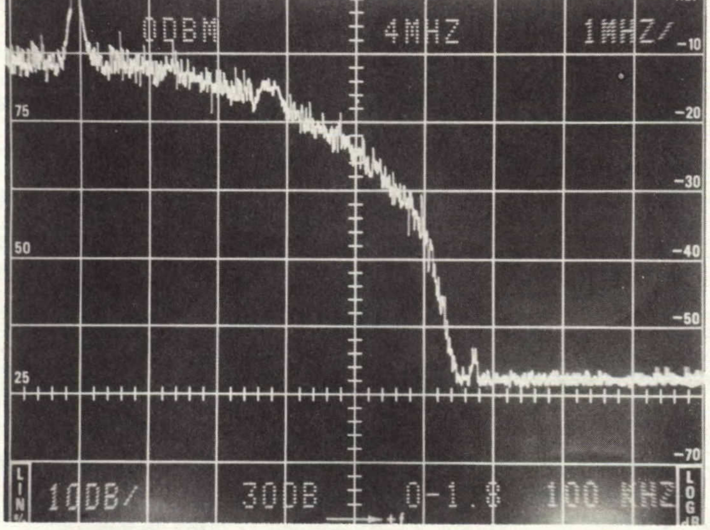


Figure 4.4: Frequency Spectrum of Pseudo-Random Sequence, with a 5 MHz low-pass filter.

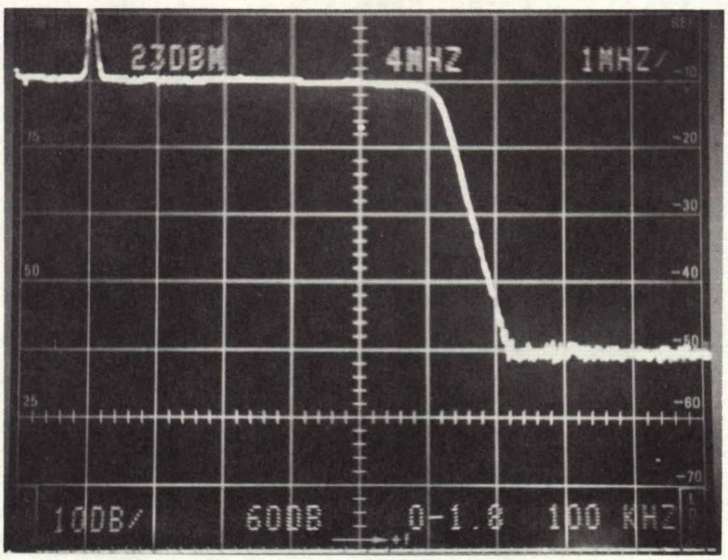


Figure 4.5: Amplitude Response of Phase-Equalized 5 MHz Low-Pass Filter.

SECTION 5. SIGNAL PROCESSING

5.1 General

After the digitization of TELETEXT lines with the LSI-11 system is completed, the data contained on the floppy disk is transferred to another computer system for analysis. This analysis is divided into two main parts: the IMPULSE RESPONSE evaluation and the TRANSFER FUNCTION computation. The results of the impulse response evaluation are used to compute the transfer function. The software performing this analysis is written in Fortran 77 and runs on a Honeywell computer system under CP-6 operating system.

5.2 Impulse Response

This first program evaluates the impulse response of the TV channel from the digitized TELETEXT data. The heart of this program is the least squares finite impulse response algorithm. Following are the details of the processing.

a) Regenerated Data

The sampled data is considered as the output of the system $y(n)$. Because the digitization is performed at two times the bit rate, the input to the system is a series of impulses (see Fig. 4.1). Therefore, the input data, $x(n)$, is regenerated from the sampled (output) data as a sequence of impulses representing zero and one levels. The "one" level is represented by the value +1 and the "zero" level by -1. The output samples $y(n)$ are offset to zero to remove the DC bias. The regeneration method avoids the need for sampling the input and the output of the system simultaneously, which is not possible with a remote broadcast system.

b) Finite Impulse Response

Next, the least squares finite impulse response algorithm is used to compute an estimate of the impulse response from regenerated (input) data and sampled (output) data. By specifying the number of points (M) to describe the impulse response the desired precision can be reached.

c) Time Window

If the transfer function is to be computed from the evaluated impulse response, a window must be applied to the impulse response to avoid in the frequency domain aliasing:

$$h_w(m) = h(m).w(m) \qquad m=0,1,...M$$

where w(m) is the sine squared (Hamming) window shown in Figure (5.1):

$$w(m) = \sin^2 \left(\frac{m\pi}{M}\right)$$

The spectrum of $h_w(m)$:

$$H_w(f) = H(f)*W(f)$$

where * denotes the convolution.

Thus, the windowing causes a widening of the spectrum of h(m).

d) Interpolation

An interpolation algorithm is also available allowing the evaluation of the impulse response at intermediate points.

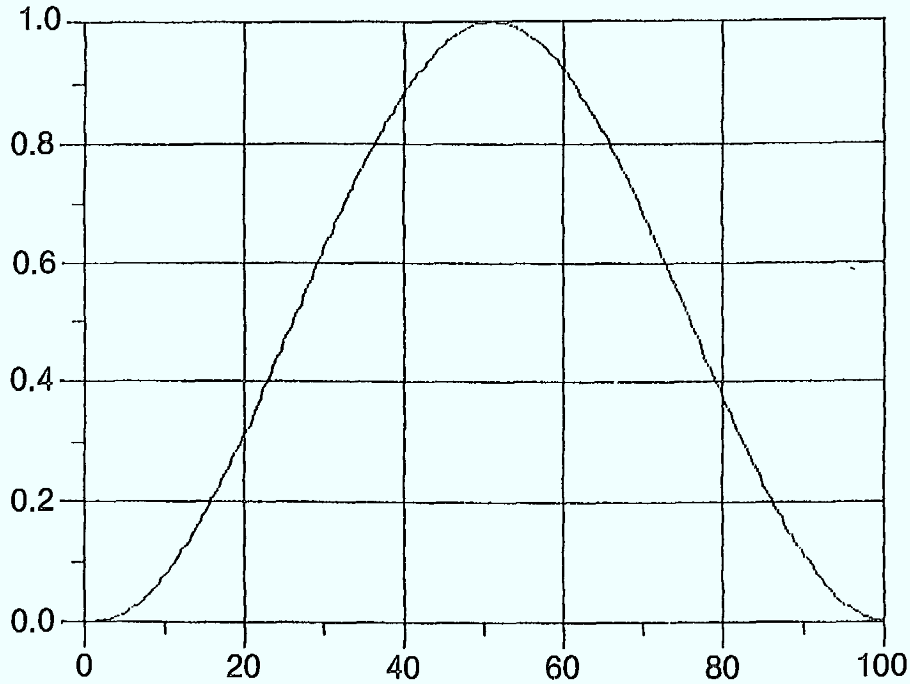


Figure 5.1: Squared Sine (Hamming) Window, M=100.

The interpolation of a sequence effectively increases the sampling rate. Therefore, from a sequence X(t) sampled at F_s, we want a sequence, Y(t), that approximates X(t) as if it would have been sampled at L^*F_s.

To do so, the algorithm starts by inserting L-1 zeros between any two samples of X(t). This increases the sampling frequency of X(t) from Fs to L^*F_s. Then, to conform with the frequency domain avoiding aliasing, the process requires to pass X(t) through a lowpass filter.

In this program, only impulse responses of a channel are interpolated. Instead of computing the frequency characteristics of the impulse response and of the lowpass filter, doing the multiplication, and, then transforming the result to the time domain, the lowpass filter is represented by its impulse response and the convolution of the two sequences is performed. To make the new sequence time finite, a raised cosine window is superimposed.

A Nyquist lowpass filter with a roll off of zero can be represented by a SIN(X)/X equation, in this interpolation process, it is:

$$\frac{\sin(n\pi/L)}{(\frac{n\pi}{L})} \qquad -ML \leq n \leq ML$$

where L is the interpolation factor

M is a function of the number of lobes of the interpolator, (M=8).

The representation of the Nyquist filter by SIN(X)/X involves a infinite number of lobes. In this processing, a number of lobes had to be used to get an acceptable approximation. Using 15 lobes (7 each side of the foundamental) combined with the raised cosine window, gives an expected error of less than 1%. The raised cosine window is given by:

$$\cos^2(\frac{n\pi}{2ML}) \qquad -ML \leq n \leq ML$$

e) Interpretation

By making a graphic plot of the impulse response, it is possible to see some interesting characteristics of the channel. In this way, echoes and filtering effect of the channel are easily seen on the curve. With some minor modifications, the complex impulse response evaluation could be implemented and thus quadrature distortion detected.

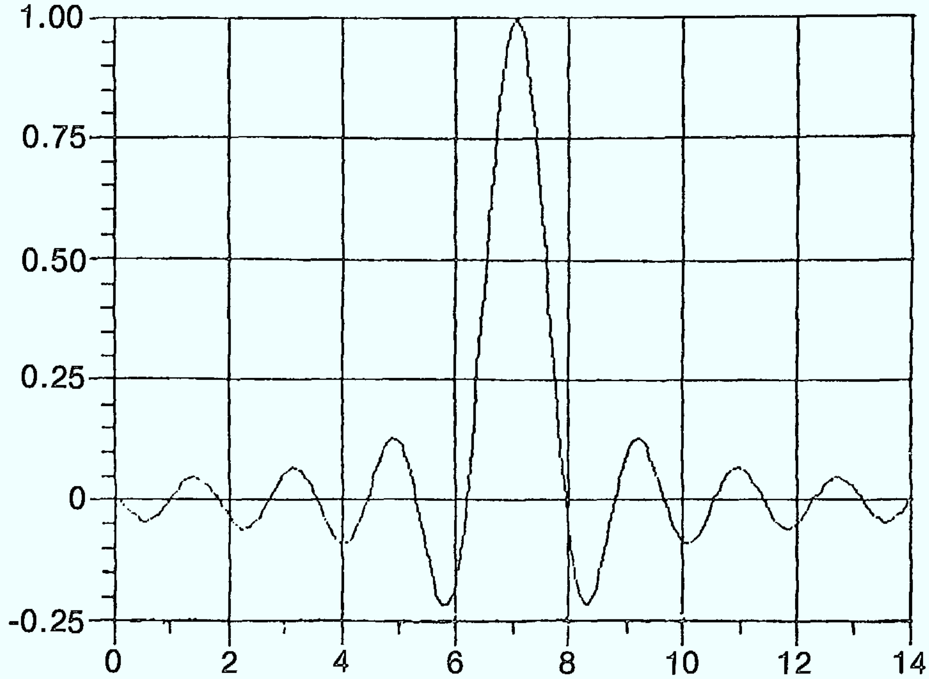


Figure 5.2: SIN(X)/X (15 lobes) Interpolator.

5.3 Transfer Function

A second program computes the transfer function of the channel from the impulse response. A fast Fourier transform subroutine is used and the amplitude and phase response are computed. Amplitude and phase data can be drawn and saved on seperate files for later processing.

a) Fast Fourier Transform

The subroutine performs a complex floating point FFT from floating point real data. As input, the subroutine requires a number of points equal to a power of two. Considering the trade off between computer limitation (memory) and the precision of the FFT, an FFT with 4096 points was chosen. The cosine and sine coefficients given by the FFT are used to compute the amplitude values and the phase values of each harmonic.

b) Amplitude and Group Delay

The amplitude and phase responses are computed directly from the sine and cosine coefficients of the FFT. The phase is then "unwrapped", because of the modulo 2 process resulting from the phase computation. The group delay is computed as the derivative of the phase with respect to frequency. Graphs of both curves can be drawn using PLOT 10 subroutines and TEKTRONIX compatible terminal.

SECTION 6. RESULTS

6.1 General

This section presents some results of impulse response measurements performed under laboratory conditions. The set-up was as shown in Figures 4.2 and 4.3. The channel was a 4 MHz low-pass filter. The teletext packets consisted of a pseudo-random sequence at a bit rate of 5.72 MB/s. The signals at the output of the pulse shaping filter, and at the output of the 4 MHz filter were digitized at a rate of 11.45 MHz. The sampling clock was derived from the colour burst signal.

Figure 6.1 shows the digitized signal at the output of the 4 MHz filter. The digitization window is such that only the RPBS is digitized.

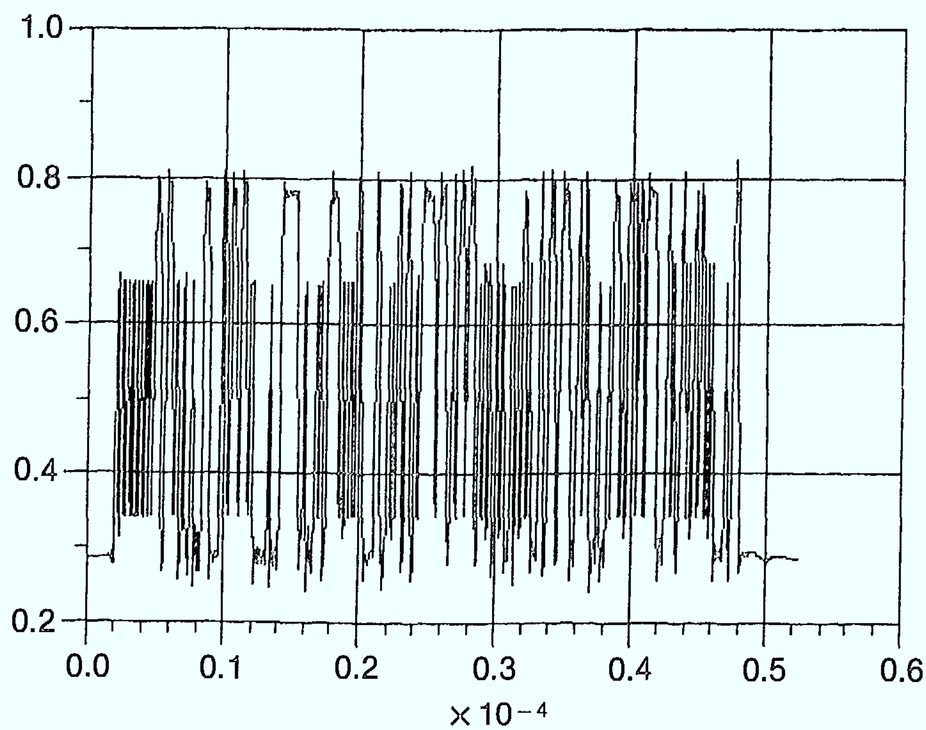


Figure 6.1: Digitized Pseudo-random Sequence at the Output of a 4 MHz LPF.

6.2 Autocorrelations

The autocorrelation functions of the regenerated input sequence x(n) (impulse train regenerated from y(n); refer to the model shown in Figure 4.1) and of the sequence y(n) at the output of the 4 MHz filter are shown in Figures 6.2 (a) and (b), respectively.

It is seen that $\psi_{xx}(k)$, the autocorrelation function of x(n), has a few side-lobes near the main lobe (k=0), due to the periodic nature of the first portion of the packets, consisting of 16 bits of alternating 1 and 0, used for bit synchronization (see Figure 6.2). This is an integral function of the Teletext encoder, and cannot be by-passed. It should be noted that $\psi xx(0)$ has a value of approximatly 0.5, due to the fact that half the samples of x(n) have a value of zero, to simulate a sampling rate of 11.45 MHz.

6.3 Pulse Responses

Two digitized signals, the output of the pulse shaping filter and the channel output, were passed separately through the FIR algorithm. Recalling that input sequence, x(n), is modelled as an impulse train, the output of the FIR algorithm is thus the system's response to a one bit (1T) pulse or the impulse response of the combination of the channel and the pulse forming filter P(w) (see Figure 4.1).

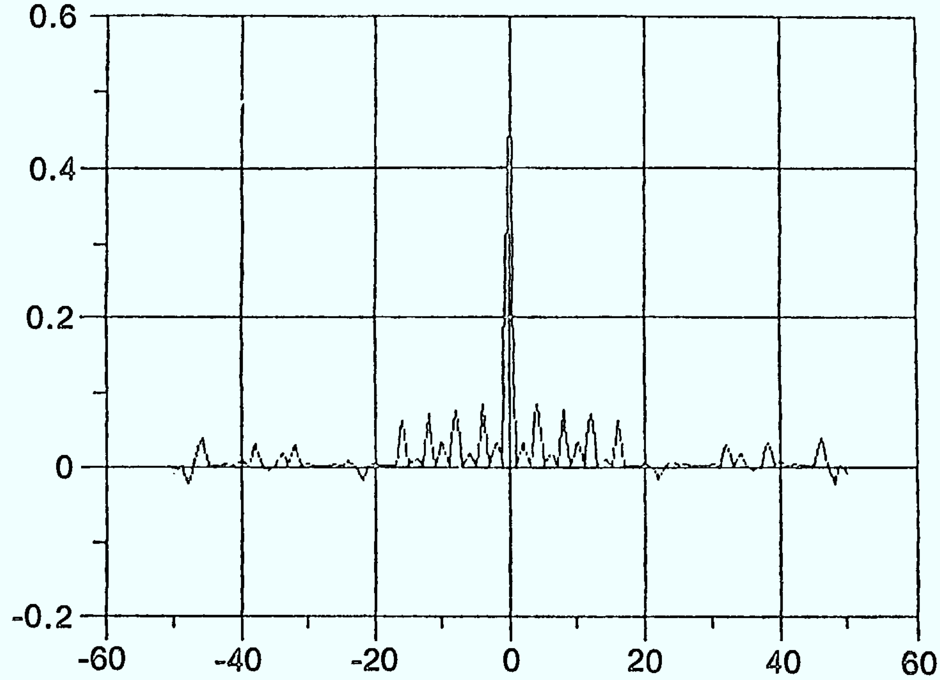


Figure 6.2 (a): Autocorrelation of the Regenerated Input x(t) (impulse train).

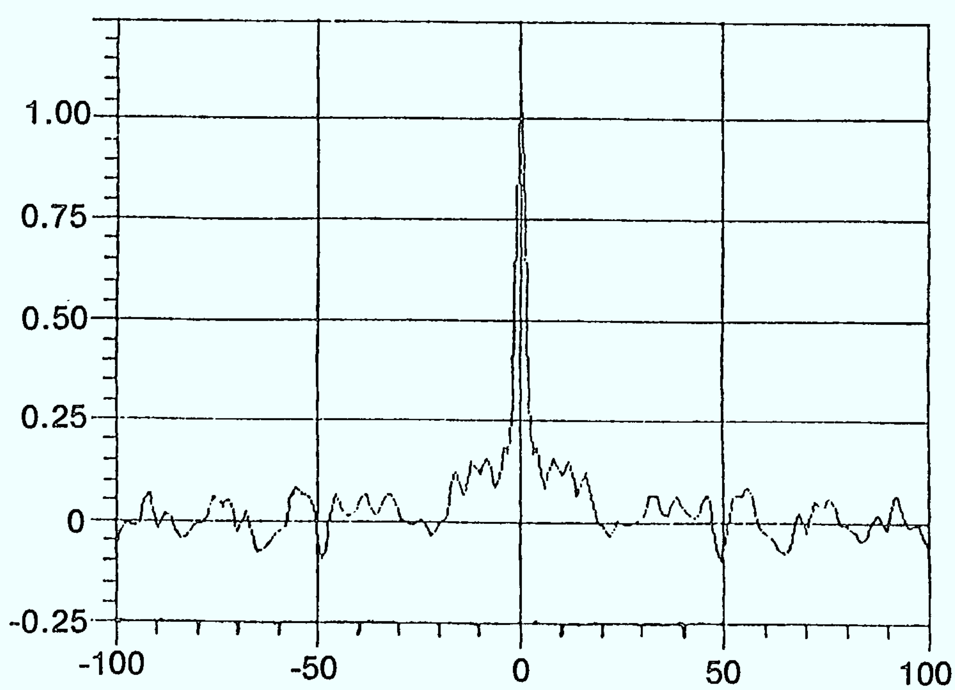


Figure 6.2 (b): Autocorrelation of the Output y(t).

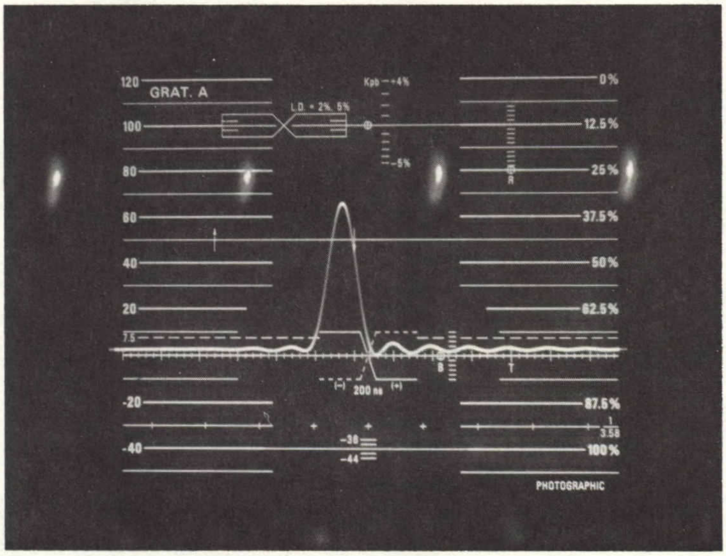


Figure 6.3 (a): Picture of 1T pulse at output of raised cosine filter

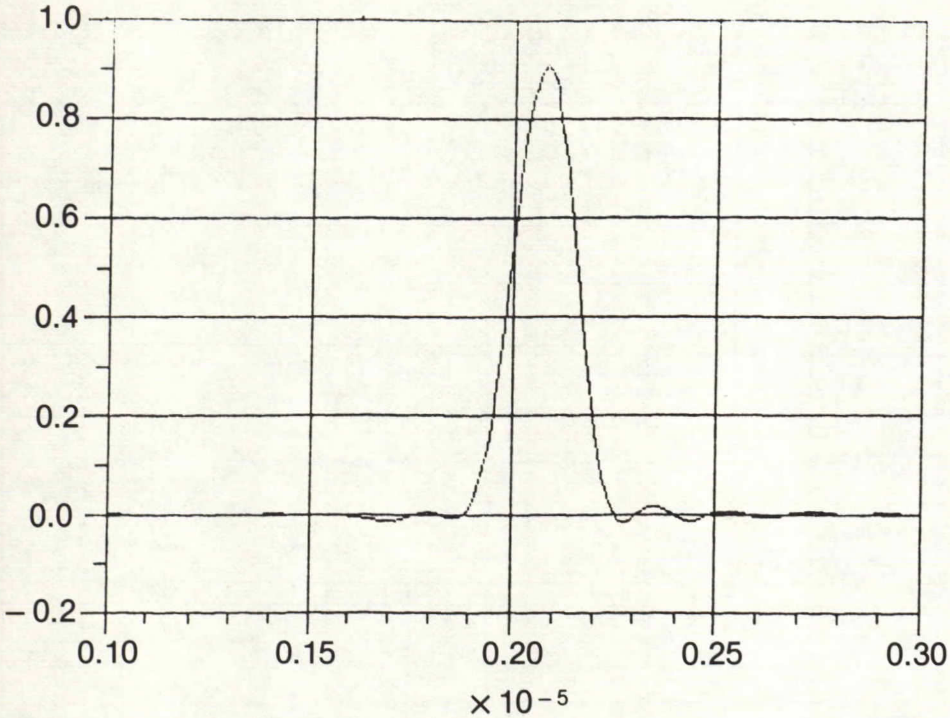


Figure 6.3 (b): 32-points estimated response interpolated by a factor of 5

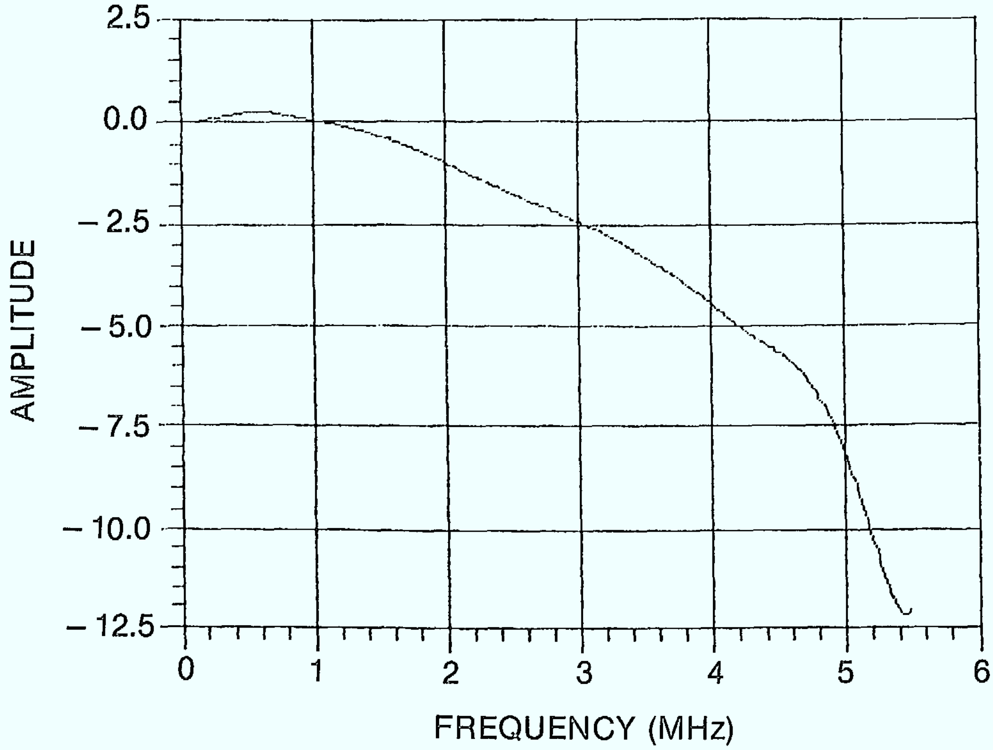


Figure 6.4 (a): Amplitude response of raised cosine filter

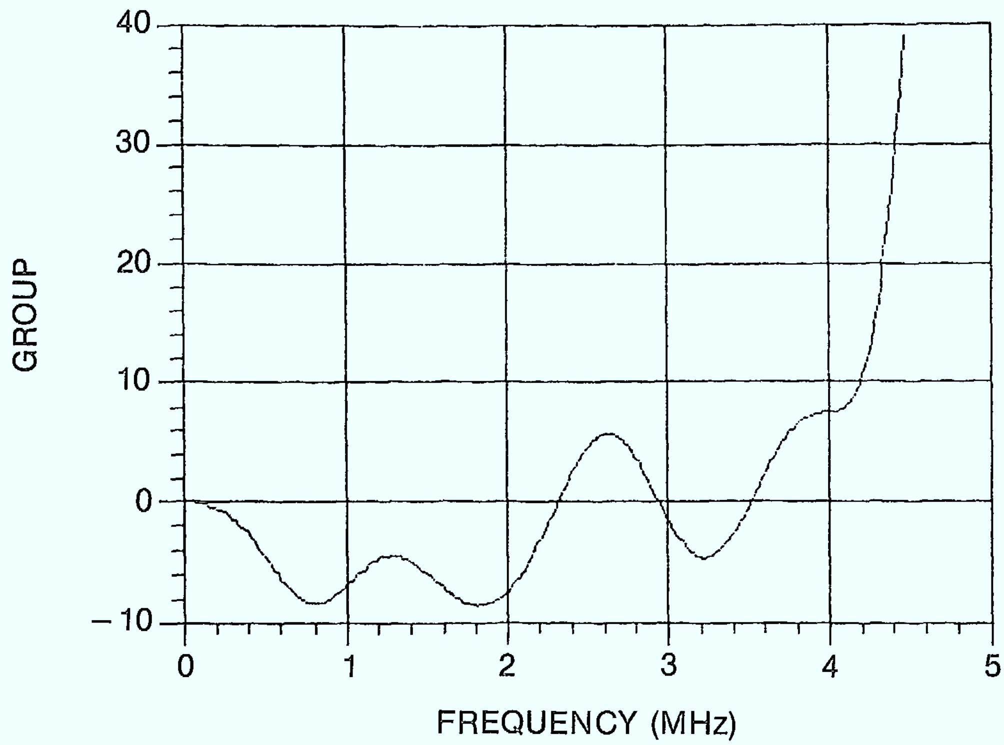


Figure 6.4 (b): Group delay (in nano-seconds) response of raised cosine filter

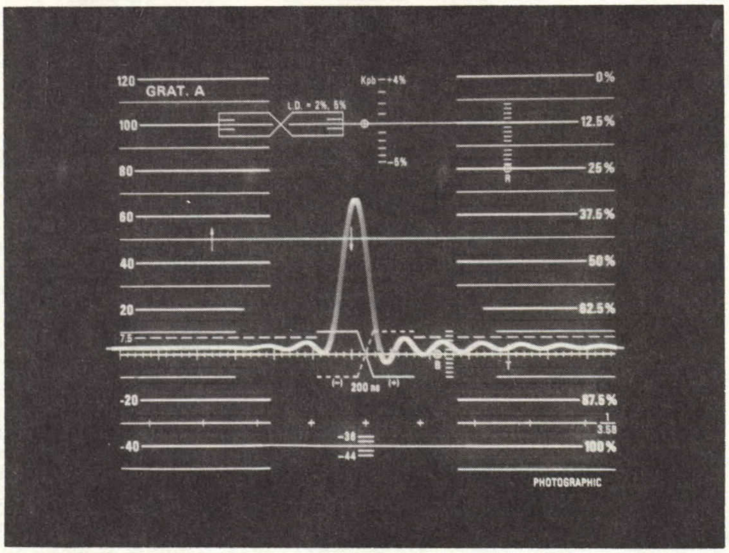


Figure 6.5 (a): Picture of 1T pulse at output of 4 MHz LPF

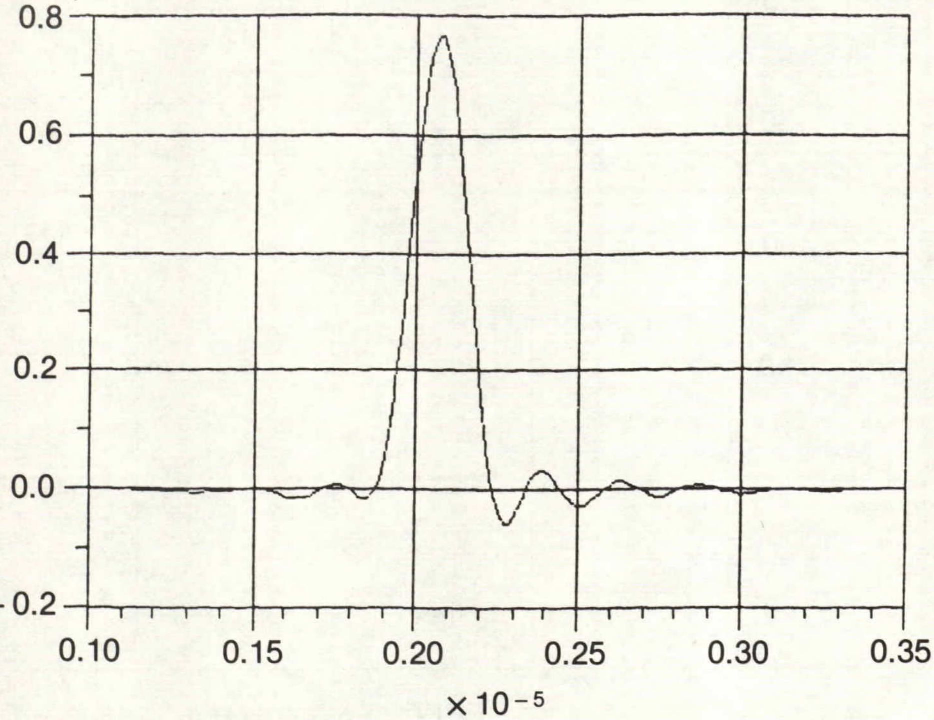


Figure 6.5 (b): 32-points estimated response, interpolated by 5

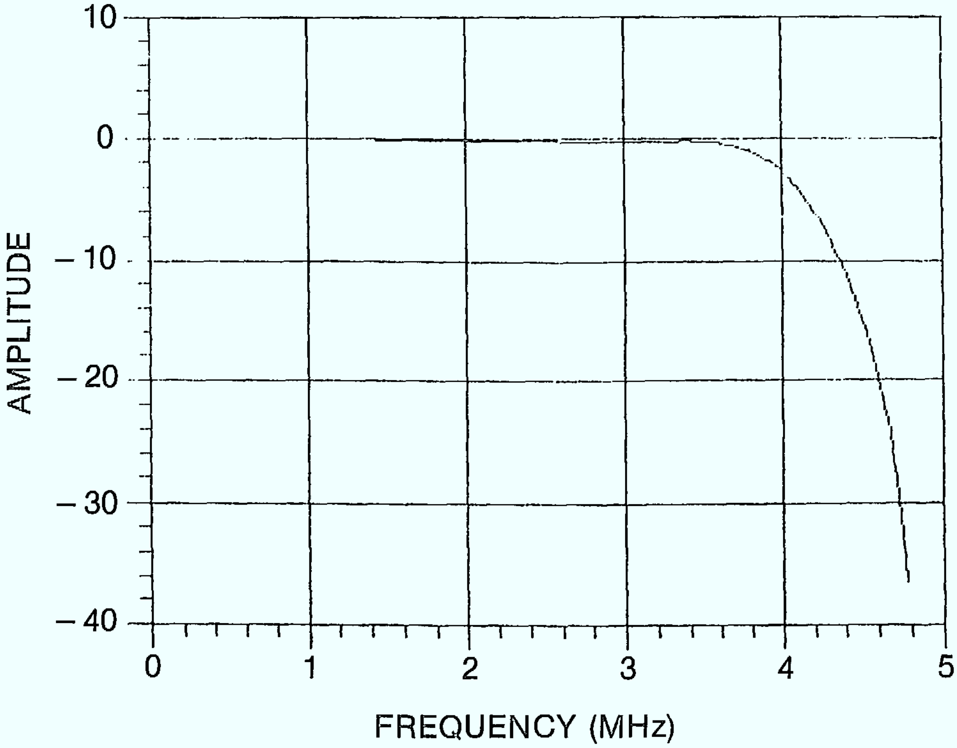


Figure 6.6 (a): Amplitude response of 4 MHz only

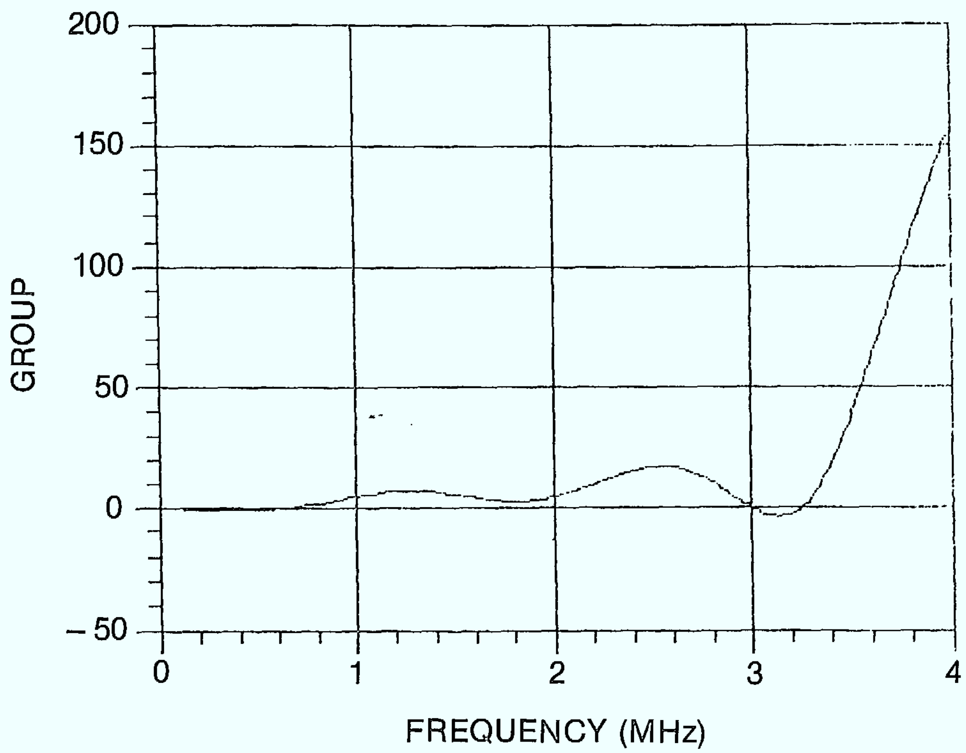


Figure 6.6 (b): Group delay (in nano-seconds) response of 4 MHz only

A 32-point impulse response was computed, for each digitized signal, and interpolated by a factor of 5, to a sampling rate of 57.272 MHz. Figure 6.3 (a) shows a picture of a 1T pulse at the output of the raised cosine pulse shaping filter, and Figure 6.3 (b) shows the estimated response: they are identical. Dividing this response by P(w), gives the impulse response of the raised cosine filter only, from which the amplitude and group delay responses are computed. These are shown in Figure 6.4 (a) and (b) respectively.

The measured and the estimated 1T pulse responses of the output of the 4 MHz LPF (which includes the pulse shaping filters) are shown in Figure 6.5. Again the results are identical. At this point, the raised cosine and 4 MHz filters transfer functions may be computed by dividing by the (sin x)/x function, or the response of the 4 MHz LPF only may be computed by dividing the Fourier transforms of 6.5 (b) with 6.3 (b) (refer to Eq. 4.2). The result is shown in Figure 6.6; it is quite close from the measured frequency response of the filter (not shown). The main difference is the width of the transition band, which is wider than in reality. This is due to the windowing of the impulse response, which causes the transition width to increase.

6.4 Noise Performance

In order to determine the estimation accuracy in the presence of noise, white Gaussian noise was added to the signal at the input to the 4 MHz low-pass filter, as shown in Figure 6.7.

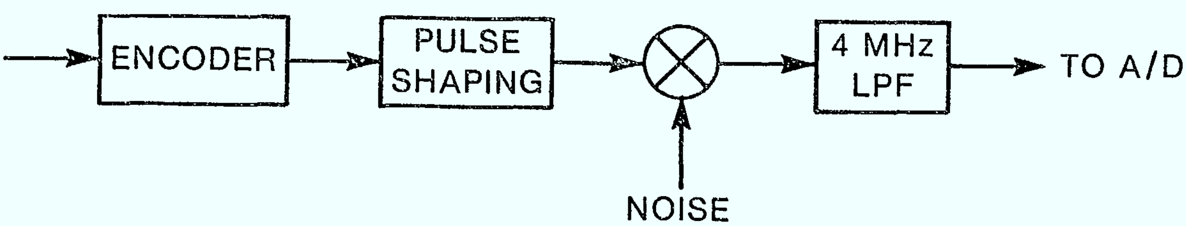


Figure 6.7: Block Diagram of System Adding White Gaussian Noise

The video signal-to-noise ratio was measured:

$$SNRV = 20 \log \left(\frac{0.714}{\sigma_n}\right)$$

where σ_n is the rms noise level.

A measure of the estimation accuracy is the squared sum error:

$$P = \sum_{n=0}^{N} |e(n)|^2$$

$$\text{where } e(n) = y(n) - \sum_{m=0}^{M} h(m).x(n-m) \qquad n=0, 1, \ldots N$$

P can be easily calculated within the FIR algorithm and is then normalized,

$$\overline{P} = \frac{1}{N} \sum_{n=0}^{N} |e(n)|^2$$

Figure 6.8 shows the variation of $\overline{P}$ with the video signal-to-noise ratio, SNRV, for this experiment. As expected, the normalized error energy increases with the noise level. At SNRV larger than 40 dB, $\overline{P}$ does not vanish due to the residual noise in the system, including quantization noise. As a comparison, Figure 6.9 shows $\overline{P}$ for a computer simulation of a low-pass filter. In this case, there is no residual noise and, indeed, $\overline{P}$ asymtotically decreases to zero.

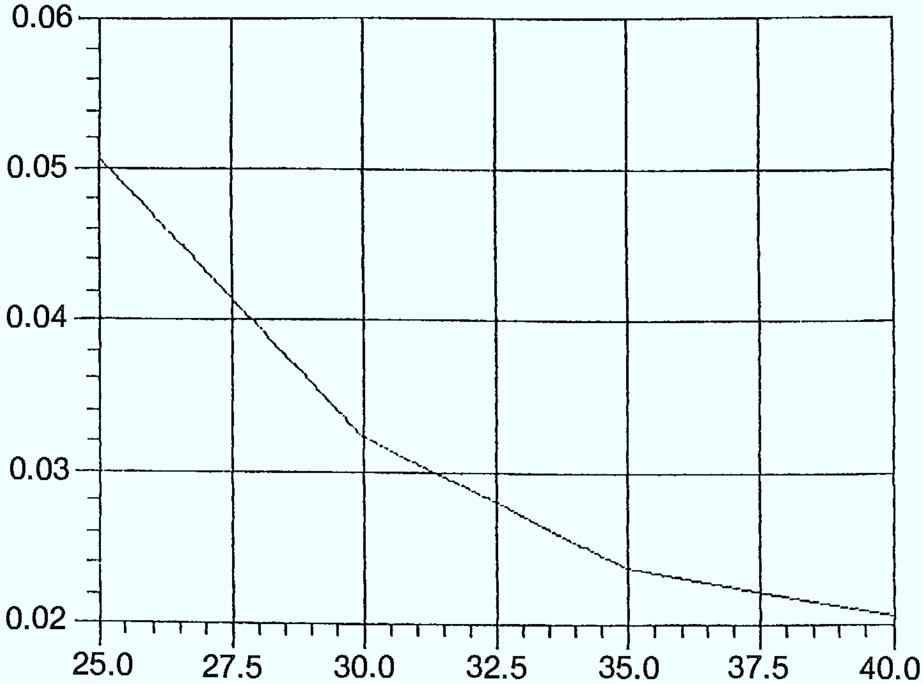


Figure 6.8: Normalized Error Energy, Versus Video SNR, for a 4 MHz LPF, 32-points Impulse Response.

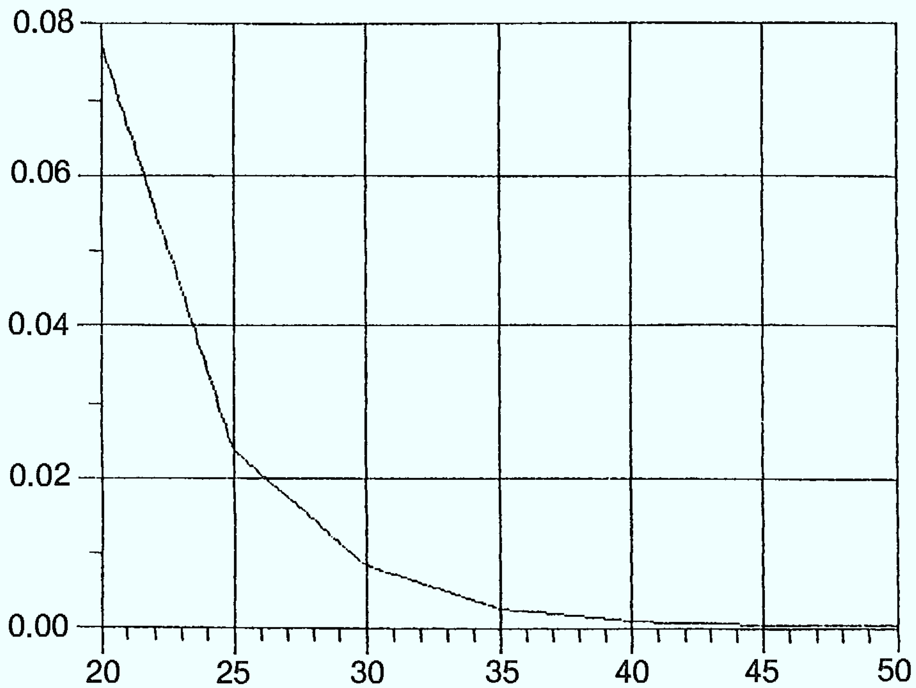


Figure 6.9: Normalized Error Energy, Versus Video SNR; Computer Simulation of Low-Pass Filter.

SECTION 7. CONCLUSIONS

A method for computing the real part of the impulse response of television channels has been described. The advantage of this system is that TV channels may be characterized, without disrupting the normal TV program, simply by digitizing the received pseudo-random signal.

This system has been used in teletext field measurements, performed by CRC. The results are quite accurate. Averaging of successive digitization improves the accuracy of the estimation, especially in a noisy environment. With the use of a quadrature demodulator, the system can be easily modified to compute the complex impulse response.

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APPENDIX

Low-Pass Equivalent Model of the Television Channel

A simplified block diagram of a television broadcast system is shown in Figure A.1. The spectrum S(f) of the transmitted signal is given by:

$$S(f) = \tfrac{1}{2}[M(f-f_c)+M(f+f_c)]T(f) \qquad \text{A.1}$$

where M(f) is the Fourier transform of the baseband signal m(t), and T(f) is the transmitter filter, shown in Figure A.2-(a). For conveniency, we shall denote $M_+(f)$ as the positive frequency content, and $M_-(f)$ as the negative frequency content:

$$S(f) = \tfrac{1}{2}[M_+(f-f_c)+M_-(f+f_c)]T(f) \qquad \text{A.2}$$

At the receiver, the signal is converted to an intermediate frequency, f_i. Assuming perfect carrier regeneration, the spectrum of v(t) is (we assume a gain of 2 in the mixer):

$$V(f) = \tfrac{1}{2}[M_+(f+f_i)T(f+f_c+f_i)+M_-(f-f_i)T(f-f_c-f_i)]$$

$$= \tfrac{1}{2}[V_-(f)+V_+(f)]$$

We note that the mixing process causes a change in the orientation of the sidebands. The signal is then filtered by the vestigal sideband filter R(f) shown in figure A.2(c):

$$Z(f) = V(f).R(f)$$

$$Z(f)=\tfrac{1}{2}[(M_+(f+f_i)T(f+f_c+f_i).R_-(f))+(M_-(f-f_i)\ T(f-f_c-f_i)R_+(f))]$$

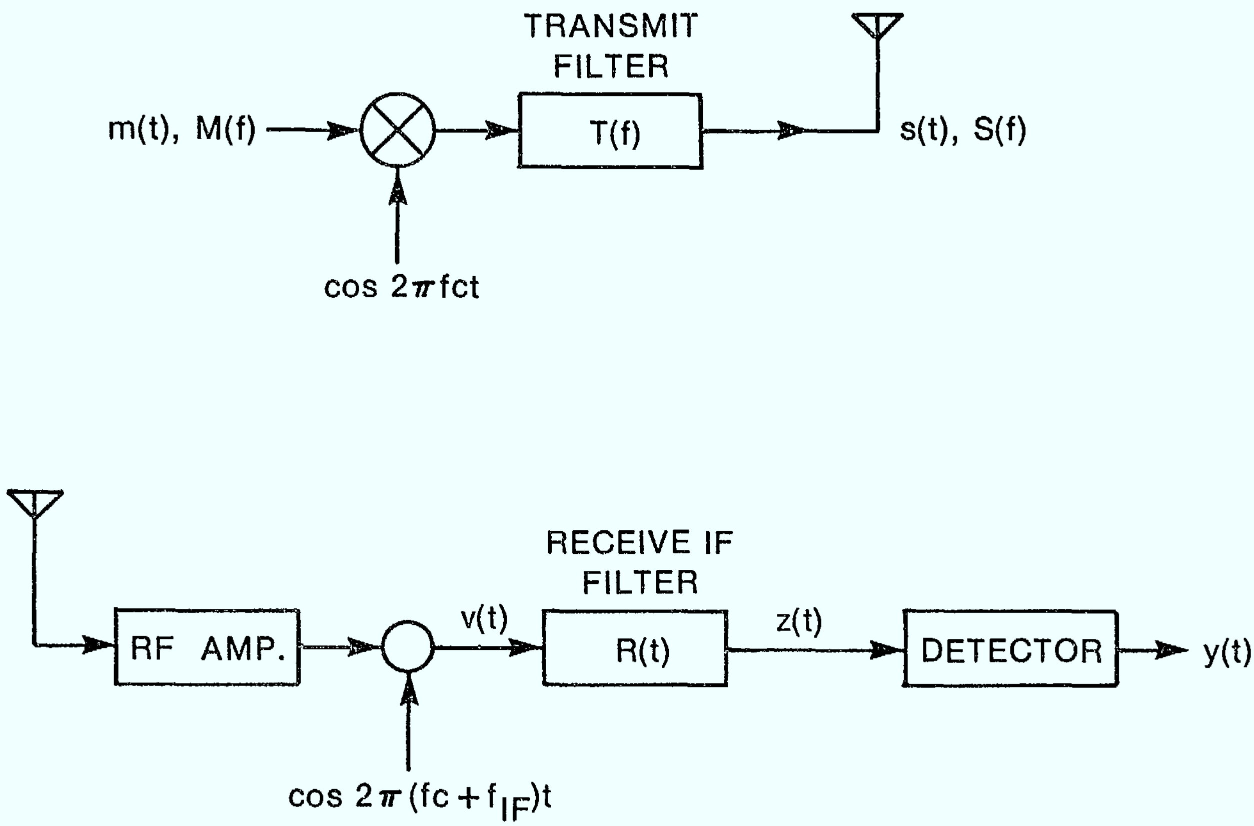

Figure A.1: Simplified Block Diagram of a Television Broadcast System.

Where $R_+(f)$ and $R_-(f)$ are respectively the positive and negative portions of R(f). This expression may be simplified by denoting H(f), the channel transfer function, as the product of the transmitter and receiver filters:

$$H(f) = T(f+f_c+f_i).R_-(f)+T(f-f_c-f_i)R_+(f)$$

$$H(f) = H_-(f) + H_+(f)$$

where $H_+(f) = H(f)$ for $f>0$
and $H_-(f) = H(f)$ for $f<0$.

The function H(f) is an unsymmetrical bandpass filter, which may be represented in terms of an in-phase and a quadrature baseband channels [8]. Lets define the two functions:

$$Hl_+(f) = H_+(f+f_i)$$
$$Hl_-(f) = H_-(f-f_i)$$

obtained by shifting $H_+(f)$ and $H_-(f)$ to the left and right respectively by f_i (figure A.4.3(a)). We also have

$Hl_+(-f) = Hl_-^*(f)$, where the sign (*) denotes the complex congugate.

The in-phase and quadrature transfer functions are given by [8] (see Figure A.4.3(b) and (c)).

$$H_I(f) = \frac{Hl_+(f)+Hl_-(f)}{2}$$

$$H_Q(f) = \frac{Hl_-(f)-Hl_+(f)}{2j}$$

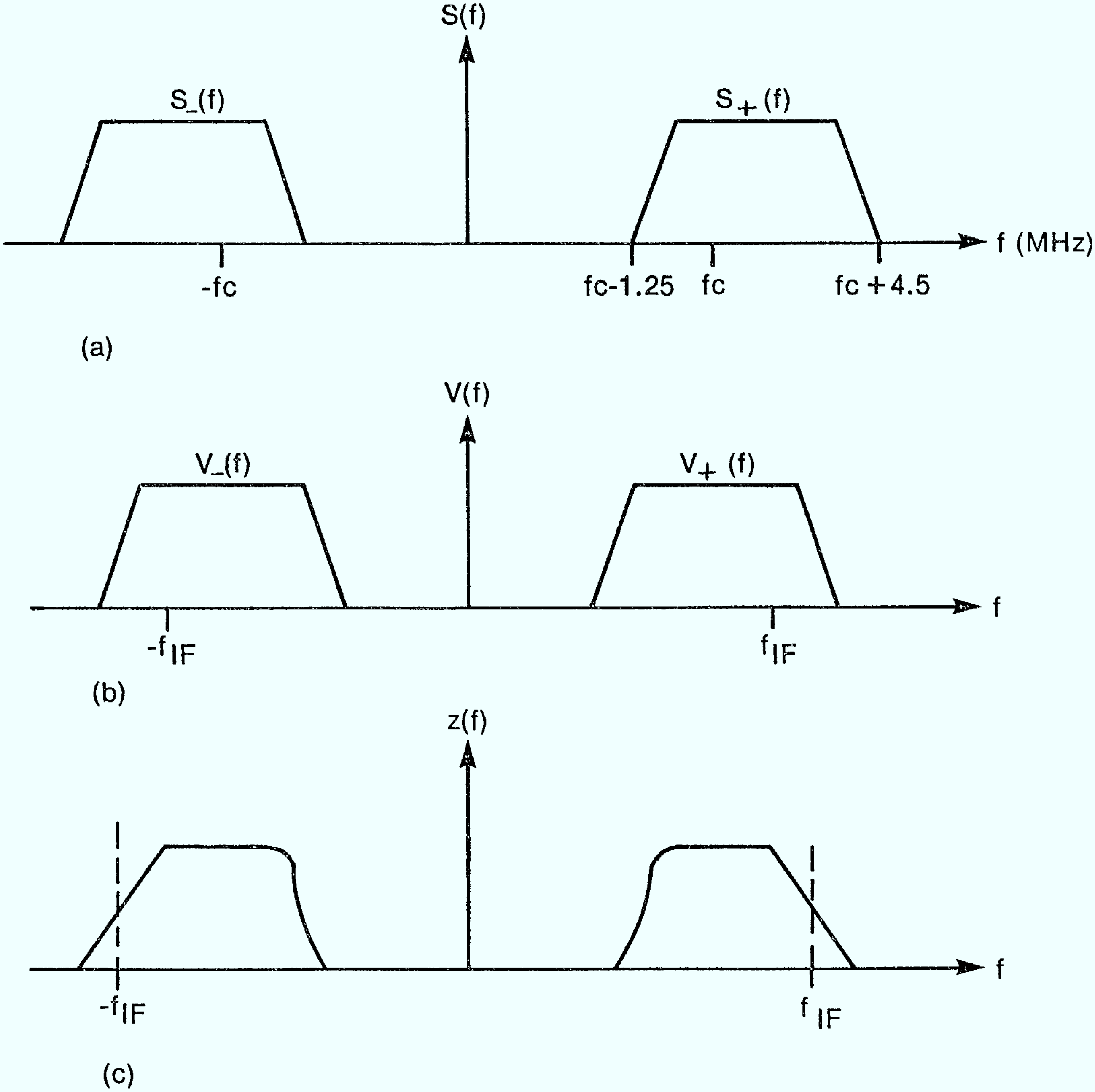

Figure A.2: Magnitude of Spectrum at the Output of the transmitter (a), after the mixer (b) and after the receiver filter (c).

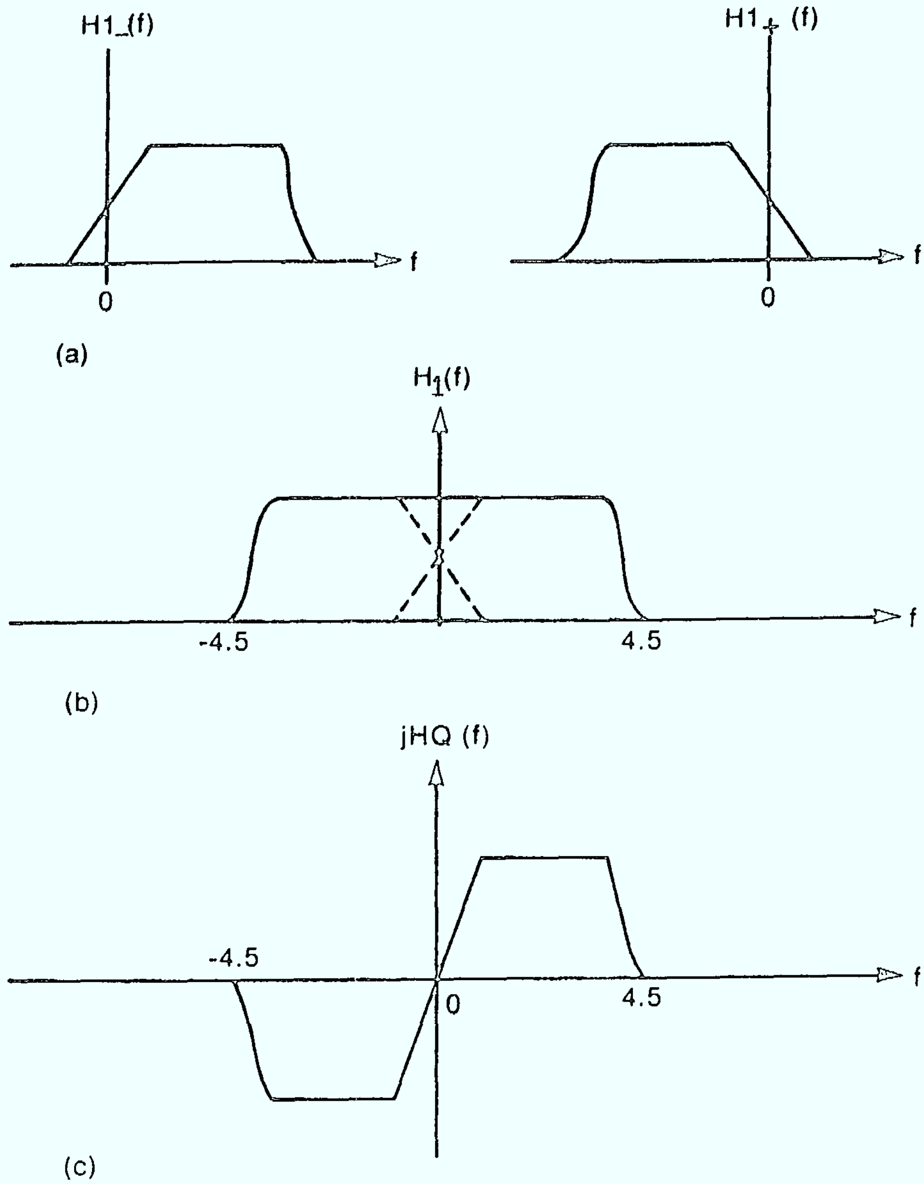

Figure A.4.3

(a) shifted + and - transfer function spectrum

(b) in-phase

(c) in-quadrature

Therefore, $H_I(f)$ and $H_Q(f)$ are Fourier transform of two real functions

$$h_I(t) \leftrightarrow H_I(f)$$
$$h_Q(t) \leftrightarrow H_Q(f)$$

The impulse response h(t) of the system H(f) can be expressed in terms of the in-phase and quadrature components:

$$h(t) = 2h_I(t)\cos(w_c t)+2h_Q(t)\sin w_c t$$

The equivalent channel model is shown in Figure A.5.

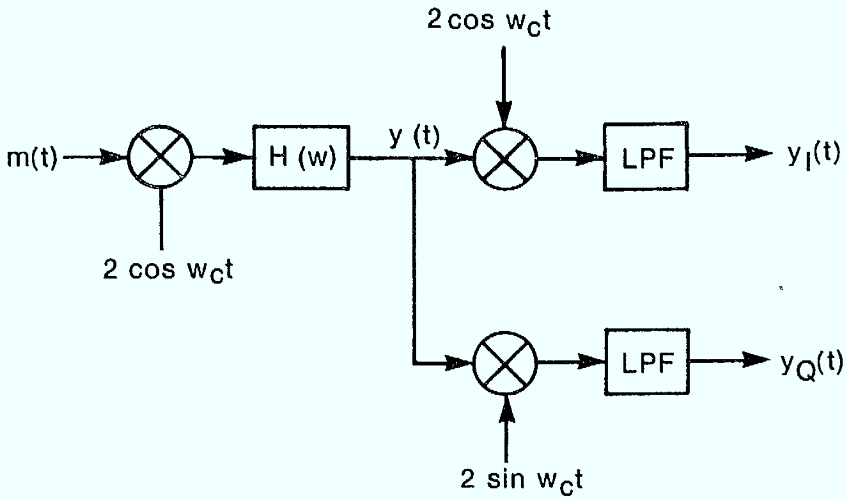

Figure A.5. Equivalent Channel Model

The system output y(t) is given by:

$$y(t)=y_I(t)\cos w_c t+y_Q(t)\sin w_c t$$

where $y_I(t) = m(t)*h_I(t)$

$y_Q(t) = m(t)*h_Q(t)$

Thus the low-pass equivalent signal is given by:

$$y(t) = y_I(t)+j\ y_Q(t)$$

and the low-pass equivalent channel model is shown in Figure A.6

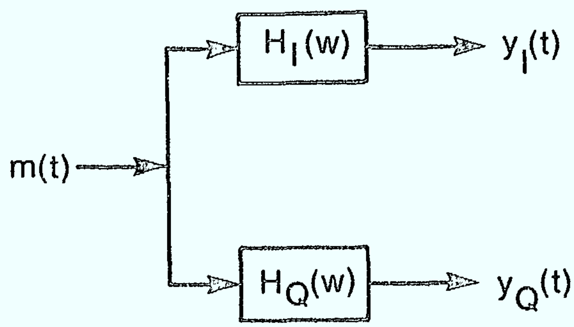

Figure A.6. Equivalent Low-Pass Model

where $H_I(w)$ and $H_Q(w)$ are the low-pass in-phase and quadrature channels respectively (see Figure A.4).

With a synchronous or quasi-synchronous TV demodulator, the received signal is multiplied by $2\cos(w_o t+\phi)$ (we assume a gain of 2); ϕ is the phase error. Under such conditions the output of the demodulator is:

$$\begin{aligned} Z(t) &= y(t)\cos(w_o t+\phi) \\ &= y_I(t)\cos\phi+y_Q(t)\sin\phi. \end{aligned}$$

There is therefore cross-talk between the I and Q channels, if ϕ is not equal to zero. If $\phi=0$, then

$$Z(t) = y_I(t)$$

The output of an envelope detector is:

$$Ze(t) = \sqrt{y_I^{2}(t) + y_Q(t)^{2}}$$

therefore, quadrature distortion is always present with this mode of detection.

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