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MULTI-DIMENSIONAL FILTER DESIGN

IN DIGITAL TELEVISION SYSTEMS.

by

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Abstract

Multi-dimensional video filtering is becoming an important part of future high-quality television systems. A digital television system is a three-dimensional sampling process, however previous work in digital video filtering has concentrated mainly on the design of one- and two-dimensional filters.

The thesis summarises the sampling and filtering techniques which may be used in video systems, and various techniques for multi-dimensional video filtering are discussed.

A new technique for flexible multi-dimensional filter design is then introduced. The new filters split multi-dimensional frequency space into sections, and these sections may then be kept or rejected to form an approximation to the desired filter characteristic. Non-separable filters of this type may be realised by a cascade of 1-D half-band filters arranged in a tree-structure, thus minimising the computation and design effort required. The number of sections in frequency space is determined by the number of decimation and interpolation stages in the filter structure and this in turn determines the resolution of the filter.

Computer simulations of the new filter structure were carried out and some of the results are included. The results are encouraging and prove that the technique

may be used to implement flexible pre and post video filters.

The thesis concludes that although the new filter structures require more hardware in real-time implementations than other dedicated filter designs, the modularity of the filter structure simplifies the hardware design (and implementation) and encourages the use of VLSI.

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Contribution to Research

The contributions to research in the fields of high-quality television systems and multi-dimensional filter design are outlined below :

i) Flexible Multi-dimensional Filter Structure:

A new multi-dimensional filter structure is introduced which has the important characteristic of realising non-separable multi-dimensional filter responses with a structure of only one-dimensional filter units. A particular implementation produces very flexible filters with no dimensional limitations and very modular hardware implementations.

The work was first published at the 1983 Picture Coding Symposium, Davis, California. The work was subsequently published in Electronics Letters, vol.19, no.25/26, 8th December, 1983.

ii) Television Filtering and Sampling Structures:

Three-dimensional filtering and sampling techniques in digital television systems are summarised and the frequency characteristics of video signals and the concept of redundancy in a picture are discussed. The new multi-dimensional filter design is introduced, and the use of the filter design in down-sampling systems and very flexible real-time adaptive systems is investigated.

iii) Subjective Effects of Video Sampling and Filtering:

A large number of subjective tests were carried out using various new filter structures and the results and implications of these tests are discussed. The filter structure has been successfully used in down-sampling video systems and some results are shown.

These results were first published at the 1983 Picture Coding Symposium, Davis, California.

NOTATION

Symbol	Definition
cpd	cycles per degree
cph	cycles per picture height
f	frequency (Hz)
f_s	sampling frequency
f_h	horizontal sampling frequency (samples/line)
f_v	vertical sampling frequency (lines/field)
f_t	temporal sampling frequency (fields/sec)
frame	one complete two-dimensional video picture (two fields in an interlaced frame)
$h(n)$	impulse response
$H(w)$	frequency response
$\|H(w)\|$	amplitude response
L	interpolation factor
M	decimation factor
N	filter order
pixel	picture element (one sample)
$P(n)$	polyphase network
T	sampling period
T_h	horizontal sampling period
T_v	vertical sampling period
T_t	temporal sampling period
$T_n(x)$	nth order Chebyshev polynomial
VLSI	Very Large Scale Integration
w	angular frequency (radians/sec)

w_S angular sampling frequency

w_S passband

w_P stopband

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CHAPTER 1

INTRODUCTION

1.1 An Introduction To Video Filter Design and Evaluation.

1.1.1 Interest in high-quality television systems.

Research into improving the present broadcast television system has been of interest to broadcasters and equipment manufacturers alike, ever since the present system was adopted. The research has fallen into three basic categories :

i) improvements in receiver design.

ii) different transmission standards, 'compatible' with present receivers.

iii) completely new systems which are not compatible with present receivers, and which represent a quantum leap in subjective quality.

European and American interest, <1,2,3>, has concentrated mainly on what Jackson and Tan <1> describe as 'compatible evolution', and consequently most of their research has fallen into the first two categories.

However, they have also researched into new systems which utilise wide-screen displays and much wider transmission channel bandwidths. Their aim has been to match the technical and subjective quality of 35mm film.

The Japanese Broadcasting Corporation <4>, on the other hand, has concentrated on the third approach and its aim is to produce a system which will eventually be adopted by the film industry. Indeed, film companies in America have already shown interest in the system.

1.1.2 Interest in bandwidth reduction and filtering techniques.

The major problem which has restricted research into high-quality systems has been transmission channel bandwidths. Terrestial transmission channels have been over subscribed for many years, and although investigations were carried out into various bandwidth reduction schemes, the work concentrated mainly on low quality video systems <5,6,7>.

However, the recent allocation of satellite channels for broadcast TV has considerably increased the level of interest in high-quality television systems. To obtain the optimum received picture quality, full use must be made of the available channel bandwidth, and any redundancy in the transmitted signal should be eliminated. These requirements, coupled with the proposed use of wide-screen displays, has led broadcasters to consider much more sophisticated methods of optimising transmission bandwidth.

1.1.3 Types of filters used at present.

In chapter 2, the present broadcast television system is shown to be an example of a two-dimensional sampling process. (The image is sampled vertically by the line scan and temporally by the field scan). It is well known that in all sampling systems, pre- and post-filtering must be employed to prevent alias and repeat spectra components appearing at the output of the system. This would seem to imply that the correct 2-D filtering should be carried out in the present broadcast television system. However, this is not the case for several reasons:

i) Correct pre-filtering at the camera is difficult to achieve. The main methods available are either optical filtering, or the careful choice of spot size and charge decay time constants. Generally, these methods only allow low-order (Gaussian) filtering.

ii) Correct post-filtering at the receiver is also difficult to achieve. Similar filtering techniques may be used, but it is still difficult to obtain more than a very basic filter.

iii) The theoretically 'correct' pre- and post-filters may not necessarily give the best perceived picture quality for a given sampling structure. For example, for certain picture material the eye may prefer some aliasing in moving scenes rather than the reduction in resolution in an alias-free picture.

A theoretical and subjective compromise must therefore be reached in designing video filters.

Due to the use of line scanning in broadcast television, the bandwidth of the video signal has generally been considered in the horizontal direction only. Filters may be realised more easily in this direction, and at present, one-dimensional analogue filters are used extensively for the separation of sound, colour and luminance components. A simple vertical filter with an analogue delay line is also used in PAL (Phase Alternate Line) receivers to decode the colour sub-carrier. These analogue filters are bulky and inflexible, and these techniques would prove impractical for multi-dimensional filter designs.

1.1.4 Problems in designing and implementing multi-dimensional filters.

1. Algebraic Factorisation.

The lack of algebraic factorisation techniques for polynomials of more than one variable complicates the design of non-separable multi-dimensional filters <8>. The amount and complexity of the computation required for the design can also prove excessive . Consequently, simpler separable filter designs have mainly been considered in the past. However, the obvious filtering limitations and lack of flexibility of these techniques are becoming more of a problem to video filter designers, especially in the subjective evaluation of processing techniques.

2. Phase Accuracy.

The phase of the Fourier transform of an image is generally more important than the magnitude in the perception of the image <9,10>. Phase accuracy in image processing is therefore extremely important. Because of this, video filters are generally of the finite impulse response (FIR) type, thus avoiding the problems of non-linear phase, stability, and finite word-length effects, associated with infinite impulse response (IIR) filters. However digital FIR filters require more storage to achieve a given magnitude response than IIR filters. This can be a problem with vertical and temporal filters, where line and field stores can be very bulky. However, digital storage is becoming cheaper and one-bit field stores are already available on a single chip.

1.1.5 Need for subjective evaluation.

In order to maximise the use of available channel bandwidth, any 'redundancy' in the transmitted video signal should be minimised or (if possible) eliminated. It is important to define and understand what is meant by the term 'redundancy' when applied to a video signal. Redundant video information is information which could be removed from the video signal without causing any perceived degradation in the displayed picture. The important word is 'perceived'. As all pictures are perceived via the human visual system, the subjective evaluation of all processing techniques is most important.

A television signal has in effect a three-dimensional bandwidth. Likewise, the human visual system (acting as a receiver) also possesses a three-dimensional bandwidth. Therefore, in theory, if the three-dimensional bandwidth of the transmitted video signal matches that of the human visual system, there will be no redundancy in the system. However, defining the bandwidth of the human visual system is very difficult. It has been shown that the bandwidth is dependent on many factors eg. :

i) the individual
ii) display contrast and brightness
iii) picture content

iv) viewing angle

v) viewing time (duration).

Much work has been carried out in the past to investigate and model the human visual system. Recent work has applied these studies to picture processing, and theoretical and practical models and yardsticks have been formulated in an attempt to formalise the results of subjective tests. However, these results are very difficult to compare because of the lack of standardisation in viewing conditions and picture material.

The present broadcast television system contains some information, such as line and field blanking, which is obviously redundant. A bit-rate saving of about 25% could be achieved if the blanking periods of the signal were utilised.

The idea of eliminating redundancy in a video signal may be extended by investigating the perceived effects of reducing the volume of the transmitted bandwidth beyond that of the human eye. Tonge <11> asks the question, "what information capacity present in a television signal can be discarded for minimum subjective picture degradation?". A possible technique for bandwidth reduction involves filtering and down-sampling the video signal before transmission and up-sampling and filtering the signal at the receiver for display. Various three-dimensional sub-sampling techniques are discussed in Chapter 2.

1.2 Review of previous work.

Initially, the previous work on sampling techniques in video systems is discussed.

The complexity of designing non-separable multi-dimensional filters, discussed earlier, has led many authors to investigate ways of extending established one-dimensional filter design techniques to two or more dimensions. The work carried out in this area, and its implications to multi-dimensional video filter design, is also covered.

Finally, the wide range of approaches to video filter design, which authors have adopted in the past, is summarised. Some of these approaches are covered more fully in Chapter 2.

1.2.1 Video sampling techniques.

The work on sampling techniques can be split into two catagories :

i) low-quality video systems

ii) high-quality video systems.

i) Low-quality video systems.

A lot of the earlier work was carried out on sampling systems for so-called 'narrow-band' or low-quality video systems, such as 'Sampledot' <12>. Techniques such as pseudorandom dot scanning were used to exploit redundancy and the high degree of correlation between adjacent pixels. Coupled with the use of low frame rates, compression ratios of up to 10:1 over the conventional system were achieved <5>. High-order line and dot interlace systems have also been investigated <6>. Line interlace factors such as 5:1 or 7:1 have given satisfactory results, with corresponding transmission bandwidth reductions of around 60%. These techniques would not prove suitable for high-quality broadcast signals, however their use for military and video conferencing applications is already established.

ii) High-quality video systems.

Three-dimensional sampling techniques for broadcast quality television systems have been researched with a view to optimising the sampling structure. Mersereau <13> has investigated the advantages of hexagonal sampling patterns for two-dimensional images, with a view to reducing machine storage and arithmetic computations. Ouellet et al. <14> and Sabatier et al. <15> have proposed and investigated non-orthogonal sampling structures for present broadcast television systems. They have chosen spatial-temporal sampling structures such as Line and Field 'Quincunx' in an effort to minimise the number of samples per unit volume of bandwidth without introducing aliasing. Recently, Wendland <2,16>, Schroder <17> and Tonge <18> have investigated orthogonal and non-orthogonal sampling structures, and have proposed decimation and interpolation schemes for future high-quality television systems.

1.2.2 Extension of one-dimensional filter design techniques.

Techniques for one-dimensional filter design are well established. Many authors have investigated the possibilities of extending these techniques to two or more dimensions. The well-known window technique may be extended to two-dimensions by techniques such as rotation <19>. The McClellan transformation <20> takes a one-dimensional zero-phase filter and uses a variable substitution method to map points in the original filter response to contours in the two-dimensional filter response. However, this method can pose problems in contour approximation. The frequency sampling method may also be extended to two dimensions. It entails formulating the two-dimensional discrete Fourier transform from a sample matrix of the frequency response and deriving the impulse response. However, the computational requirements of this method, especially if linear programming methods are used, may be restrictive.

Filters may also be designed using multi-dimensional projections <21>. For example, the filtering of a three-dimensional signal by a two-dimensional filter operating in a given plane, can be described in the frequency domain as a multiplication of the projection of the spectrum of the signal in the corresponding plane by the frequency response of the filter.

1.2.3 Video filter design.

Many areas of work in picture processing have stimulated research into filter design, such as :

Bandwidth reduction
Low- and high-quality sampling systems
Noise reduction
Luminance/chrominance separation
Standards conversion.

Depending on the application, the solutions to these problems have ranged from simple analogue filters to complex adaptive digital filters.

Relatively simple non-linear (adaptive) temporal filters have been investigated by Dennis <22> for noise reduction in video conference coders. These filters are considered to be non-linear because the spatial filter characteristic depends on the amount of movement in the picture in a non-linear fashion. Crawford <23>

implemented an adaptive spatio-temporal filter for prefiltering in video conference coders. Here the input signal is fed to a two-dimensional spatial filter and a one-dimensional temporal filter, and a movement detector determines the proportion of each output signal used in forming the final output. Clarke <24> and Parker <25> have implemented more sophisticated non-linear vertical/temporal 'comb' filters for improving the luminance and chrominance separation in PAL television pictures. Here, the 'direction' of the comb filter (in two-dimensional space) is determined by a movement detector. Tanaka et al. <26> and Achiha et al. <27> have implemented three-dimensional filter structures for use in line-scan standards conversion. Again the filter response is obtained by using a movement detector to change the coefficients of a spatial filter in a non-linear fashion.

Wendland <16> has designed linear FIR planar filters for pre- and post-filtering 'offset' (diagonal) sampling structures. Mersereau <13> has applied the window and transform techniques, described earlier, to the design of two-dimensional FIR filters for use with hexagonal sampling patterns.

Ouellet and Dubois <14> have designed three-dimensional filters for use with non-orthogonal sampling structures. One-dimensional filters, designed by the method of projections, were combined to form the final filter structure.

Drewery <28> and Tonge <29> have investigated the design of linear three-dimensional FIR filters by specifying the response of the filter along cross-sections in three-dimensional frequency space. Drewery applied the filters to luminance/chrominance separation on PAL signals, and Tonge concentrated on pre- and post filtering for various sampling patterns.

1.3 Outline of the Thesis.

Following on from the brief introduction in this chapter, the idea of sampling and filtering a video signal in three-dimensions is discussed in Chapter 2. The importance of understanding the human visual system is emphasised and the idea of eliminating redundancy in a video signal is introduced. The matching of the sampling structure to the response of the eye is important, and a variety of possible sampling structures is investigated. The design of pre- and post-filters for each of these sampling structures is then discussed. A summary of design techniques for two- and three-dimensional filters is also given.

In Chapter 3, a new technique for designing flexible non-separable multi-dimensional filters is introduced. The new technique may be realised by arranging one-dimensional half-band filters in a tree-structure, and therefore the design of filters is greatly simplified. Examples of some possible implementations, such as two- and three-dimensional tree-structures, are presented. Finally, a theoretical analysis of the design technique is given.

Chapter 4 describes the computer simulations which have been carried out to investigate the new filter structure. The simulation software and half-band filter designs are described and a summary of the results is given.

Conclusions on the research, and suggestions for further work, are given in chapter 5.

Appendix 1 contains a brief summary of the computer simulation software.

CHAPTER 2

SAMPLING AND FILTERING IN DIGITAL TELEVISION SYSTEMS.

2.1 Introduction.

The importance of the human visual system in investigating image processing techniques can not be over estimated. Therefore the chapter begins by describing the use of human visual models in picture processing, and discussing some important characteristics of the eye. Some methods of exploiting the response of the visual system are mentioned.

Chapter 1 mentioned that a television system is a sampling process. One-, two- and three-dimensional sampling processes in television systems are discussed and some examples of decimation and interpolation systems are presented. To make full use of the available transmission channel bandwidth, the video sampling structure should be chosen carefully. In fact, it is desirable that the sampling structure supports a signal bandwidth which matches that of the eye. Various sampling structures are shown, together with their associated pre- and post-filters.

Finally, a summary of two- and three-dimensional video filter design techniques is given.

2.2 The Human Visual System.

The study of human vision has be recognised to be of central importance in the design of image processing systems <30>. Image processing techniques are requiring increasingly complex human visual models so as to maximise their effectiveness. Many visual models have been developed <31,32,33,34,35,36> and although some of these models are quite elaborate, major differences still exist between the behaviours that these models can emulate and the actual behaviour of the human visual system <37>. Granrath <37>, mentions the eye's ability to adapt to a wide variety of ambient scene conditions as a possible explanation for this.

In the specific area of television, researchers have carried out subjective tests on many picture processing techniques such as sampling and pre- and post-filtering <38,39,40>. However, the results of these tests are very difficult to compare due to the lack of standardisation in picture material and viewing conditions.

Chapter 1 mentioned that the response of the eye is dependent upon many factors, including picture content, viewing angle, and display parameters such as contrast and brightness. In fact, our visual system can adapt to scene illumination intensities spanning nine orders of magnitude <41> and a scene contrast of three orders of magnitude <42>. The spatio-temporal transfer functions of the human eye, as developed by Budrikis <31> and Koenderink et al. <32>, are shown in fig.2.1. Both authors give similar results. The response for small changes in intensity (luminance, fig.2.1a,b) is 'bandpass' both spatially and temporally. For small changes in colour (chrominance, fig.2.1c), the response is 'lowpass' both spatially and temporally. These responses highlight a number of characteristics of the human visual system which are already exploited by television ;

i) In both cases (luminance and chrominance), the response of the eye falls to zero at around 30Hz. It is because of this property that television systems may successfully reproduce moving images with only 25 or 30 frames per second.

ii) In both cases, the spatial bandwidth of the eye gradually reduces as the temporal frequency is increased. Section 2.4 shows that the line-interlace scanning used in present television systems exploits this characteristic to reduce the transmission bandwidth.

However, these models do not tell the full story.

The response of the eye is also dependent on orientation <43,44>. In fact the eye's bandwidth for 45 degree diagonal frequencies can be 20% down on its response for horizontal or vertical frequencies. The response is also dependent on intensity. For example, the spatial and temporal cut-off frequencies when looking at a bright CRT (cathode-ray-tube) display are around 60cpd (cycles per degree) and 70Hz respectively <43>. If the intensity is reduced by a factor of 10,000 the cut-off frequencies reduce to about 6cpd and 25Hz.

In conclusion, the response of the eye must be taken into account in order to optimise the use of the available transmission bandwidth.

2.3 Television as a Sampling Process.

Fig.2.2 shows that the present 'analogue' television system is an example of a two-dimensional sampling process. The scene is sampled in the vertical direction by line-scanning, and in the temporal direction by field-scanning. In the PAL television system used in the U.K., these sampling frequencies are 625cph vertically, and 50Hz temporally. These two scanning mechanisms allow an essentially three-dimensional image (spatial and temporal information) to be represented and transmitted by a one-dimensional signal. If the signal is sampled for digital processing, the image is effectively sampled horizontally to form individual picture elements (pixels). The resulting sampling pattern is three-dimensional and supports a three-dimensional bandwidth.

It is well-known that sampling a one-dimensional signal with a sampling frequency of :

$$w_S = 2\pi/T \quad , \text{ where } T = \text{sampling period} \qquad \ldots.(2.1)$$

gives rise to an infinite number of repeats of the baseband spectrum in the frequency domain. The repeat spectra are centred at integer multiples of the sampling frequency. The Nyquist theorem states that if the original signal is bandlimited to within the range

$$-\pi/T < w < \pi/T \qquad \ldots.(2.2)$$

then the repeat spectra will not overlap. The signal can therefore be reconstructed (without error) with the appropriate post-filtering.

This idea is easily extended to three-dimensions where sampling periods of T_h (horizontally), T_V (vertically) and T_t (temporally), give rise to repeat spectra centered at integer multiples of

$$2\pi/T_h \ , \quad 2\pi/T_V \ , \quad 2\pi/T_t$$

respectively. For example, the simple orthorombic sampling lattice in the time domain, shown in fig.2.3a, gives rise to the pattern of repeat spectra in the frequency domain shown in fig.2.3b. The Nyquist bandwidth that this sampling pattern can support without aliasing is now represented by a volume in three-dimensional space and is shown for positive frequencies by the shaded solid. Cross-sections through the solid for T=0, H=0 and V=0 are shown in fig.2.3c. In fact the bandwidth may be represented by any solid which tessellates on the repeat spectra structure. For accurate reconstruction, the video signal should therefore be pre- and post-filtered with filters of this three-dimensional shape.

Tonge <18> has found that three-dimensional filter theory has many analogies in the field of crystal diffraction and he has introduced some of the terminology of crystallography into video filter design. The sampling lattice in the time domain generates a 'reciprocal lattice' (denoting the repeat spectra centres) in the frequency domain. The 'first Brillouin zone' of the sampling lattice is in fact the primitive cell of the reciprocal lattice and therefore defines the frequency limits (or 'primitive bandwidth') of the baseband signal.

The overall 'sampling rate' for the video signal is given by :

$$f_s = f_h \cdot f_v \cdot f_t \qquad \text{....(2.3)}$$

where f_h = samples/line

f_v = lines/field

f_t = fields/second

The present television scanning system, represented by fig.2.2, is not orthogonal for two reasons :

i) The line-scan is interlaced from field-to-field (section 2.4.2).

ii) The scanning structure is lexicographic <45>. That is, it is 'tilted' due to the fact that the end of each line corresponds in time to the start of the next line, and the end of each field corresponds in time to the start of the next field.

For the sake of simplicity, scanning tilt will be neglected in the sampling pattern diagrams, however Tonge <18> has shown that the theoretical effects of this scanning tilt are negligible.

It is also important to realise that the scanning mechanism shown does not generate an infinite sampling lattice in the horizontal and vertical directions. However, these so-called 'window' effects are also negligible.

2.4 Orthogonal and Non-orthogonal Sampling Structures.

2.4.1 Orthogonal Sampling

An example of an orthogonal sampling structure has been shown in fig.2.3. This would be the sampling structure generated by what is known as a 'sequential' scanning system. However, the discussion in section 2.2 concluded that the frequency response of the human visual system was far from orthogonal, and therefore this would seem to suggest that an orthogonal sampling structure would not prove optimum.

2.4.2 Line-interlace Sampling

The interlace scanning structure used in present television systems may be a more appropriate scanning structure. Interlace scanning may be treated as a down-sampled version of the sequential scanning structure, where the number of lines in each field has been halved in an off-set fashion, fig.2.4a. This is effectively a down-sampling operation along the diagonal in the vertical/temporal plane. This generates extra repeat spectra in the frequency domain (fig.2.4b) which halve the diagonal bandwidth supported in the vertical/temporal plane, fig.2.4c. This in effect means that the vertical resolution in a scene is at a maximum when the scene is stationary, and gradually reduces as the scene begins to move. The earlier discussion of the human visual system would seem to imply that this reduction in vertical resolution is consistent with the spatio-temporal bandwidth of the eye, and therefore will not be perceived. In practice, a slight degradation in vertical resolution may be noticed, depending on the picture material and viewing conditions. However a 2:1 reduction in transmission bandwidth (over a sequential system) has been achieved, and this scanning system is still used extensively in broadcast television.

It is interesting to note that the correct three-dimensional filtering shown in fig.2.4 is not carried out in the present television system, and this gives rise to errors in the displayed picture such as the visibility of the line-structure and flicker.

2.4.3 Field-Quincunx Sampling.

A logical extension of reducing the vertical bandwidth for moving pictures (line-interlace) is to reduce the horizontal bandwidth for moving pictures as well. Down-sampling in the horizontal-temporal direction has the effect of generating extra repeat spectra along the horizontal-temporal frequency diagonal and so reducing the bandwidth along that diagonal. A sampling structure which achieves this is 'field-quincunx'.

Fig.2.5a shows the sampling lattice for three-dimensional field-quincunx sampling.

The lattice has been generated from the line-interlace sampling lattice by omitting alternate horizontal samples in an interleaved fashion. This has the effect of down-sampling the signal along the horizontal/temporal and horizontal/vertical diagonals. The extra repeat spectra generated by the process of horizontal down-sampling are shown in fig.2.5b. The subsequent loss in horizontal resolution for moving pictures and diagonal spatial resolution is shown in fig.2.5c. The volume of the primitive bandwidth is half that for line-interlace and so the transmission bandwidth has also been halved.

The shape of the primitive bandwidth for field-quincunx sampling closely resembles the approximate three-dimensional bandwidth of the human eye presented earlier. The important similarities are :

i) The spatial bandwidth gradually decreases with with an increase in temporal frequency (movement).

ii) The resolution for diagonal spatial frequencies is less than that for pure horizontal or vertical frequencies.

2.5 The Design of Two- and Three-dimensional Filters.

Existing design techniques for two- and three-dimensional filters are summarised in this section. Where appropriate, the advantages and disadvantages of these methods, with respect to video processing, are discussed and examples of their use are given.

As an introduction to the problems of multi-dimensional filter design, a fundamental problem associated with the design of multi-dimensional filters is now explained :

The design of non-separable multi-dimensional filters is complex. An underlying reason for this complexity is that the fundamental theorem of algebra, in one dimension, does not extend to two dimensions. Huang et al. <8> have described this as a "fundamental curse". In one dimension, any polynomial P(z) of degree 'n', may be factorised into n first-degree factors :

$$P(z) = K(z-Z_1)(z-Z_2)(z-Z_3)\ldots\ldots.(z-Z_n) \qquad \ldots.(2.4)$$

where K and Z_i are constants. No corresponding factorisation exists for a polynomial in two or more independent variables. It is therefore difficult to examine the stability of these filters or to realise the filters as cascade or parallel structures.

2.5.1 Two-dimensional filters.

Numerous techniques have been developed for designing two-dimensional filters <19,46,47>. Because of the problems associated with two-dimemsional mathematics, many of these methods utilise established one-dimensional filter design techniques.

1. Windowing :

The well-known window technique in one-dimension may be easily extended to two-dimensions. For example, Huang <48> has shown that a good symmetrical one-dimensional window, $w(x)$, may be used to obtain a good circularly symmetric two-dimensional window, $w_2(x,y)$, by the conversion

$$w_2(x,y) = w(\sqrt{x^2 + y^2}). \qquad ....(2.5)$$

Separable windows of the form

$$w(x,y) = w_a(x).w_b(y) \qquad ....(2.6)$$

may also be used in certain cases.

Mersereau <13> has developed window formulations for planar filters in hexagonal sampling systems

Designs based on the window method may be obtained relatively quickly, however the filters do not generally approximate to the ideal frequency response as well as other designs.

2. Frequency sampling and linear programming :

As in the one-dimensional case, the two-dimensional filter may be specified by giving values to the filter's Discrete Fourier Transform (DFT) at certain frequencies (eg. in the passband and stopband regions), and leaving the DFT samples at other frequencies (eg. in the transition band region) as variables. The error in the approximation is then minimised by varying the unspecified DFT samples. For optimum filters, (in the Chebyshev sense) all the samples are allowed to vary and linear programming methods may be used for the minimisation <47>.

The main problem with these methods is the relatively long computational time that is often required.

3. Transformations :

A two-dimensional zero-phase FIR filter may be designed from a one-dimensional zero-phase filter by using a transformation developed by McClellan <20>.

The frequency response of the one-dimensional zero-phase filter, with impulse response h(n), is given by :

$$H(w) = \sum_{n=0}^{N} a(n)\cos nw \qquad \ldots.(2.7)$$

where $a(n) = \{ h(0) \;;\; n=0$

$\{ 2h(n) \;;\; n>0$

Using the fact that cos(nw) may be expressed as an nth degree polynomial in cos(w), the frequency response may be rewritten as:

$$H(w) = \sum_{n=0}^{N} a(n).Tn[\cos(w)] \qquad ....(2.8)$$

where Tn[x] is the nth order Chebyshev polynomial.

The McClellan Transformation is then effectively the substitution of cos(w) by a two-dimensional function :

$$\cos(w) \dashrightarrow F(w_1, w_2). \qquad ....(2.9)$$

Thus for every value of w there is a contour on the w_1, w_2 plane, whose shape is determined by the function $F(w_1, w_2)$. The complete filter design therefore requires the design of the one-dimensional filter and the choice of a two-dimensional contour function.

Mersereau et al. <49> and Mecklenbrauker et al. <50> have investigated this technique in depth. The advantage with this method is that large filters may be designed relatively quickly.

Mersereau <13> has also developed a two-dimensional contour function for planar filters in hexagonal sampling systems.

4. Projections :

Mersereau and Dudgeon <51> have developed a technique for mapping two-dimensional arrays to one-dimensional arrays where columns of a two-dimensional array are concatenated to form a one-dimensional sequence. A two-dimensional filter may be designed from an ideal frequency response by:

i) concatenating one-dimensional 'slices' of the required two-dimensional frequency response,
ii) designing a one-dimensional filter which approximates to the concatenated frequency response, and
iii) back-projecting the unit sample response of this filter to give a two-dimensional unit sample response.

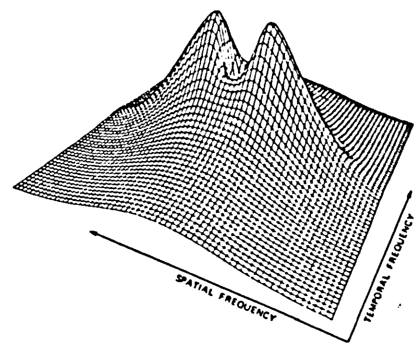
2.5.2 Three-dimensional filters.

Some of the design techniques discussed in the previous section could possibly be extended to three-dimensions. However, the amount and complexity of the computation required would be prohibitive, especially for the design of flexible filters. Therefore, three-dimensional filter designs have tended to use a mixture of one- and two-dimensional designs, in various filter structures, to obtain an approximation to the required three-dimensional response.

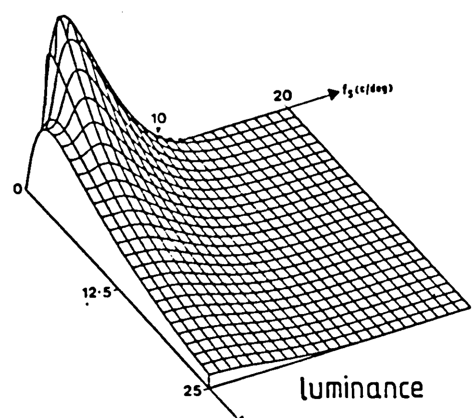
In the design of separable three-dimensional filters for use in NTSC television signals, Ouellet and Dubois <14> have designed three one-dimensional filters using the method of projections. The filters act in two different directions in three-dimensional frequency space, and are combined to form the required three-dimensional response.

In the design of non-separable three-dimensional video filters, Drewery <28> and Tonge <29> have used the idea of determining one-dimensional cross-sections through a three-dimensional frequency response. Although Tonge has not used the method of projections directly as discussed in the previous section, coefficients in the three-dimensional filters were derived by specifying filter contours in various directions in three-dimensional frequency space.

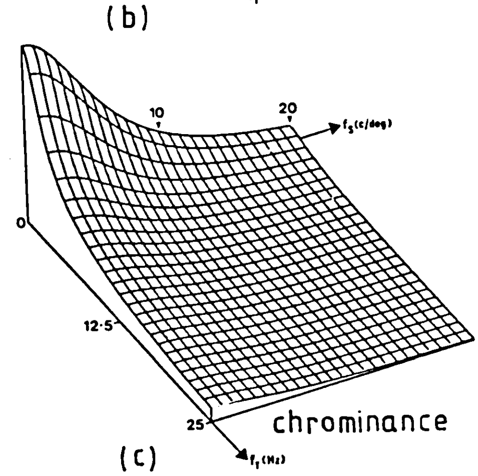
Although these design techniques have proved satisfactory in specific cases, they do not lend themselves to the realisation of a very flexible three-dimensional filter structure. In the next chapter a new approach is introduced which has proved to be both flexible and relatively easy to implement.



(a) BUDRIKIS



(b)



(c)

(b) & (c) KOENDERINK

Fig. 2.1 Spatial-temporal frequency response of the eye.

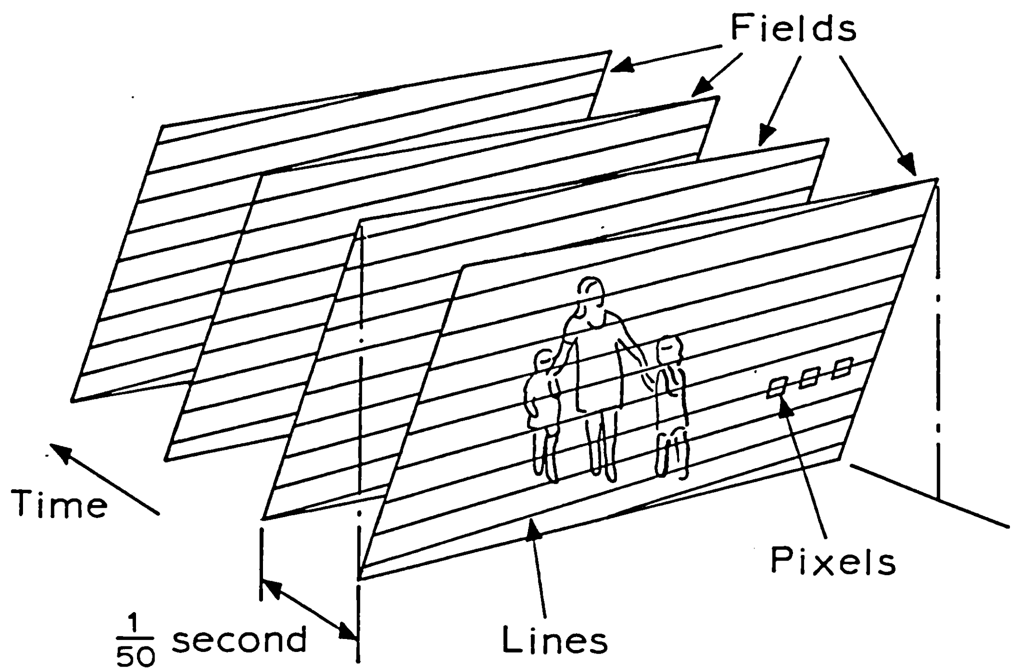


Fig. 2.2 Scanning structure for broadcast television.

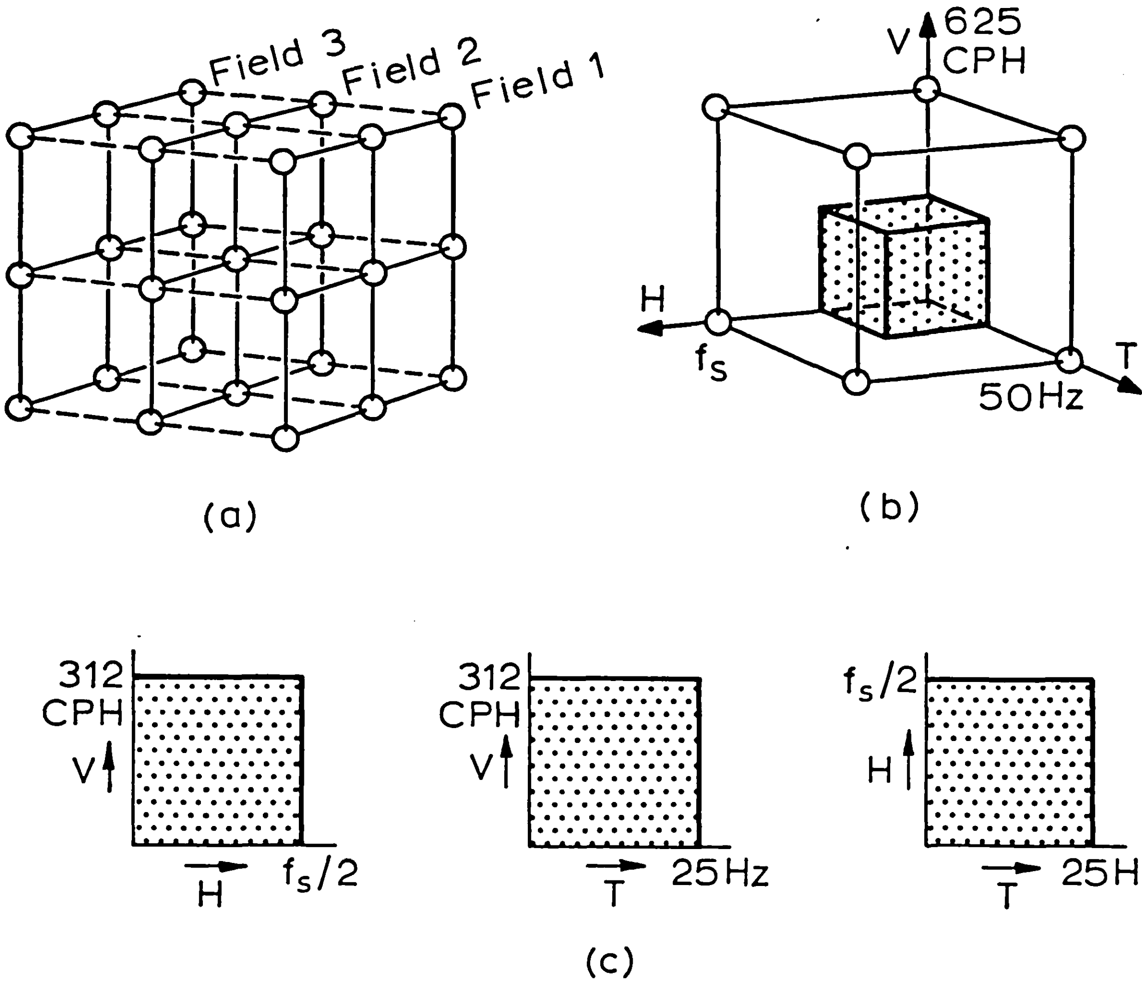


Fig. 2.3 3-D orthogonal sampling.

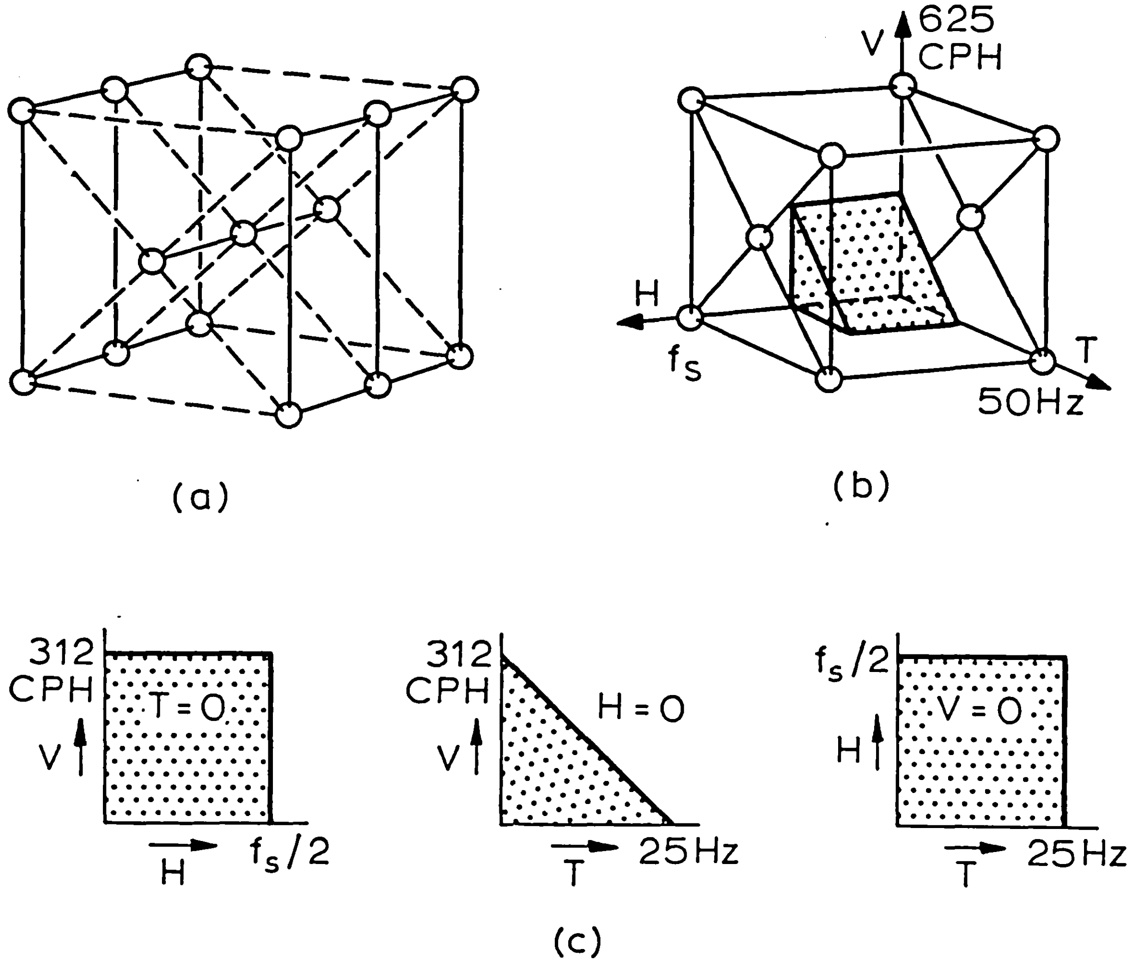


Fig. 2.4 3-D line interlace sampling.

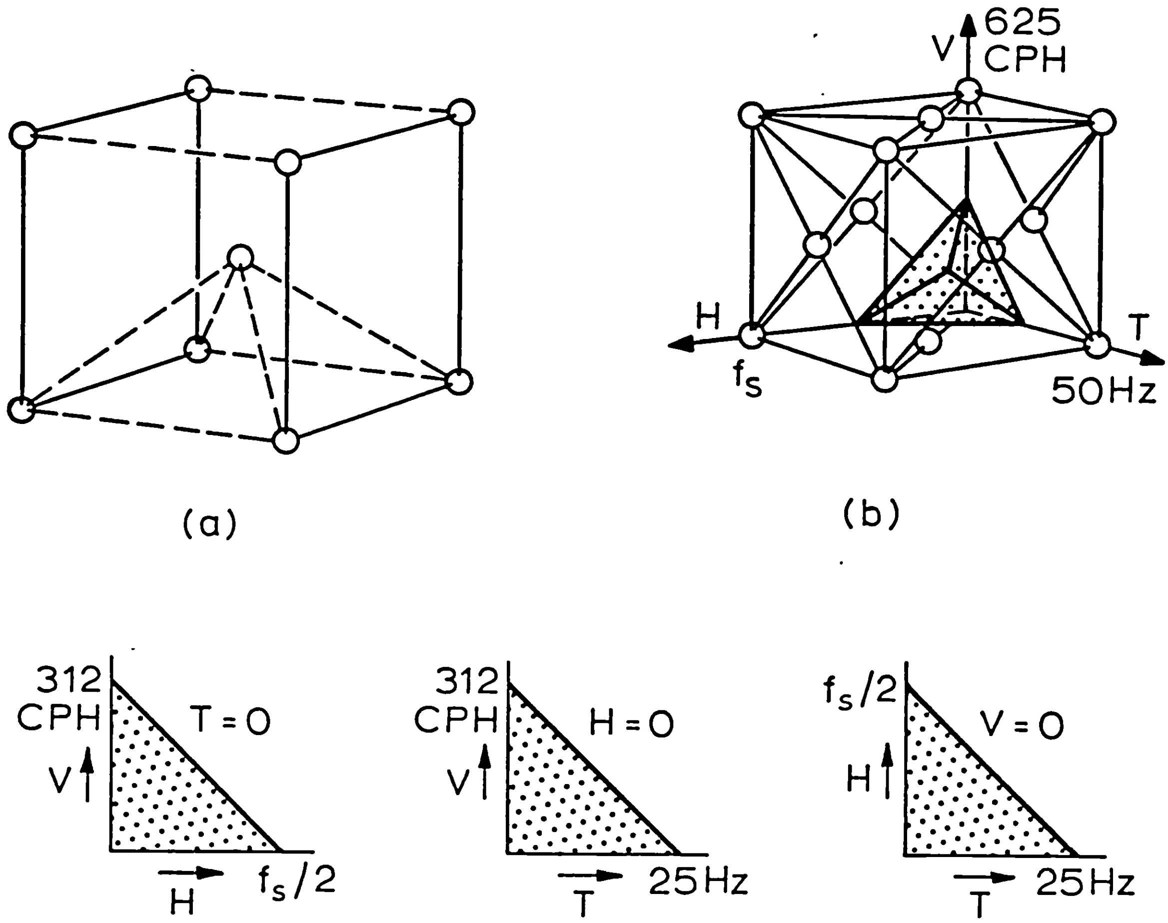


Fig. 2.5 3-D field quincunx.

CHAPTER 3

A TECHNIQUE FOR FLEXIBLE FILTER DESIGN.

3.1 Introduction.

Chapter 2 emphasised that subjective evaluation is necessary if sampling structures and filter characteristics are to be optimised in television systems. Very few of the three-dimensional filter designs mentioned in section 2.5.2 have the flexibility to enable different sampling and filtering techniques to be assessed with the same digital hardware. They do not readily lend themselves to control by computers or microprocessors because of the complexity involved in designing and building flexible real-time hardware, which may have to operate with sampling frequencies in excess of 50MHz.

If possible, the filter design technique should have the following characteristics :

i) The filter design should utilise established design techniques to minimise the design complexity of the complete filter.

ii) The filters should lend themselves to real-time adaption very easily. For example, they should be easily controlled by a movement detector or a similar real-time input.

iii) The filter structures should avoid large real-time computations so as to minimise the hardware speed requirements.

iv) The filter structure should be modular and therefore lend itself to VLSI.

In this chapter, a new filter synthesis method is introduced which actually meets all the above criteria. The fundamental ideas behind the synthesis method are described, and examples of hardware implementations are given. Decimation and interpolation techniques are discussed briefly and some methods of implementing these operations are presented. Finally, a theoretical analysis of a tree-structure implementation utilising half-band filters is given.

3.2 A New Multi-Dimensional Filter Synthesis Technique.

3.2.1 Frequency Domain Representation.

A one-dimensional signal may be split easily into two separate components in frequency space by, for example, using lowpass and highpass horizontal filters (fig.3.1a). These individual components may be split up further in the frequency domain by, for example, using further lowpass/highpass/bandpass filters (fig.3.1c). Examples of these filters applied to a 625-line television system are also shown in figs.3.1b,c. If all the individual frequency components were made available at the outputs of the filters, selective combinations of the frequency components would produce different 'filtered' versions of the output signal, for example fig.3.2a. The splitting of the input signal frequency components represents a decimation process, and the combination of the individual frequency components at the output of the filters represents an interpolation process. The resolution of the filter output, and therefore the ability of the structure to approximate to any given filter characteristic, is dependent upon the number of decimation stages. If the decimation process is carried on ad infinitum, the final interpolation proccess is, in theory, capable of reproducing any filtered version of the input signal. Figure 3.2b gives an example of where the filter resolution has been increased to eight components in the horizontal direction.

These decimation/interpolation structures may easily be extended to two dimensions. One-dimensional filters acting in the horizontal direction are used to split the frequency response into horizontal components, and then one-dimensional filters acting in the vertical direction are used on each of the horizontal components to produce all the required two-dimensional frequency components (fig.3.3a). Again, more one-dimensional filters may be used to increase the resolution of the filter (fig.3.3b). In this way, one-dimensional filters have been used to synthesis two-dimensional filtering operations. For example, an approximation to the planar filter required for field-quincunx sampling in two-dimensions (section 2.4.3) is shown in fig.3.4.

There is no reason why these interpolation-decimation structures may not be extended to three (or more) dimensions with the advantage that only one-dimensional filters need be designed (fig.3.5).

3.2.2 <u>The Use of Half-band Filters.</u>

Complete modularity in the decimation and interpolation structures may be achieved by splitting each frequency component at the appropriate centre frequency. This implies decimation by a factor of two at each stage. This may be achieved very efficiently with the use of one-dimensional half-band non-recursive filters. The frequency response of a non-recursive half-band filter is given by :

$$H(w) = h(0) + 2\sum_{n=1}^{M} h(2n-1).\cos\{(2n-1)w/T\} \qquad ....(3.1)$$

where 1/T is the sampling frequency.

The frequency response of a typical non-recursive half-band filter is shown in fig.3.28a. An important property of non-recursive half-band filters is that all non-zero even index coefficients are null and this represents a significant hardware reduction over other filters.

Decimation by a factor of two produces an even filter resolution which is not biased towards any particular filter characteristic. However, the frequency components do not necessarily need to be split at the centre frequency. If, for example, one-dimensional filters which split the frequency components in the proportion 1/4 : 3/4 (fig.3.6a,b) are used, the final filter structure would be biased towards filters with low cutoff frequencies (fig.3.6c,d).

3.3 Tree-structure Implementations.

The previous section has shown that multi-dimensional filter characteristics may be approximated by a cascade of one-dimensional half-band non-recursive filters. A highly modular way of arranging the half-band filters is in the form of a tree-structure. Decimation and interpolation tree structures are not new. For example, Constantinides et al. <52> and Tsuda <53> have discussed the advantages of tree-structure implementations for decimation and interpolation filters in the design of transmultiplexers.

A tree-structure implementation for the one-dimensional filter of fig.3.1c is shown in fig.3.7. The half-band filter units form a symmetrical tree-structure, with the first half of the filter performing the decimation process, and the second half performing the interpolation process. The first stage splits the incoming signal into lowpass and highpass components. The two filters which form the second stage of the decimation structure act on each of the lowpass and highpass components, separately, to form the four components shown in fig.3.1c. These components may then be kept or rejected by the switches at the centre of the filter structure. The third and fourth stages of the filter perform the interpolation process and combine the chosen frequency components to form the final filter output. In fact, the interpolation structure is shown to be the circuit dual of the decimation structure (3.6.1).

To allow the general specifications to be the same for each filter unit, and therefore to maintain modularity, each filter unit should be acting on baseband signals. In the decimation process this implies that the sampling rate will be halved at each stage of the filter structure. For example, fig.3.8 shows the effect of using full-rate half-band filters on a signal. Fig.3.9 shows the effect of halving the sampling rate at the outputs of the filters. The effect has been to 'modulate' the high-pass component down to a 'baseband' position. Halving the data rate at each unit also means that the total data-rate at the output of each stage is constant, which minimises the hardware and timing requirements.

Tree-structure implementations for the two- and three-dimensional filter structures of figs.3.3a and 3.5 are given in figs.3.10 and 3.11 respectively. Again, the structures are symmetrical about their centre, the first half performing the decimation process and the second half performing the interpolation process. The operation of these two structures is very similar to that of the one-dimensional structure just described. For example, the six stages in the three-dimensional structure operate as follows :

Stage 1 - The horizontal half-band decimator unit splits the incoming signal spectrum into horizontal lowpass and horizontal highpass components. Both components have one-half of the input sampling rate (fig.3.1b).

Stage 2 - The two vertical half-band decimator units act on each of the horizontal components to give four separate two-dimensional frequency components. All output components have one-quarter of the original sampling rate (fig.3.3a).

Stage 3 - Four separate temporal half-band decimator units split each component into temporal lowpass and highpass components to form eight separate three-dimensional frequency components at the centre of the filter. Each component has one-eighth of the input sampling rate (fig.3.5).

These eight components may then be kept or rejected by the eight switches, depending on the required filter characteristic. For example, the position of the switches in fig.3.12a will give the filter characteristic shown in fig.3.12b, which is a crude approximation to the characteristic required for field-quincunx sampling in three-dimensions.

Stage 4 - The four temporal interpolator units interpolate (combine) the chosen frequency components in the temporal direction. The resulting four components have one-quarter of the input sampling rate each.

Stage 5 - The two vertical interpolator units interpolate the four frequency components in the vertical direction. The resulting two components have one-half of the input sampling frequency each.

Stage 6 - The final interpolator unit interpolates the two components in the horizontal direction to form the final filter output at the original input sampling frequency.

In practice, a more efficient implementation of the three-dimensional structure may be made, (fig.3.13). Temporal filters require field stores for the filter delay elements, whereas vertical filters require line stores and horizontal filters require only pixel stores. It is therefore more efficient if the filter units are arranged so that the units requiring the least amount of hardware are at the centre of the filter structure. The temporal decimator and interpolation units have therefore been placed first and last in the structure to minimise storage hardware.

3.4 Advantages and Disadvantages of the New Technique.

Chapter 5 suggests alternative implementations of the filter technique for future study. However, the advantages of the implementation described in the previous section are numerous. In this section the advantages and disadvantages of designing and implementing the tree-structure filter are discussed.

3.4.1 Design Advantages.

i) The decimation and interpolation units may be designed easily using any of the established one-dimensional filter design techniques, such as windowing or frequency sampling. Numerous computer programs are available for all these techniques and the use of a particular computer programme which employs linear programming techniques is described in Chapter 4.

ii) The specification of the complete filter characteristic is made simpler by the block structure in the frequency domain.

iii) The complete filter characteristic may be changed very easily and very quickly by gating the data paths at the centre of the filter. The complication of computing and down-loading different multi-dimensional coefficient matrices (encountered in many previous designs) is eliminated. Real-time adaption methods, such as the use of movement detectors or the use of different filter characteristics for varying picture content, is thus simplified considerably.

iv) Complex filter characteristics are very useful in studying the human visual system, and they may be designed relatively easily using this technique.

3.4.2 Hardware Advantages.

i) The hardware structure is modular which aids hardware design and lends the structure to possible VLSI implementation.

ii) Hardware and computational requirements have been reduced through the use of half-band filters. Requirements may be reduced even further by implementing half-band filters designs with coefficients restricted to sums of powers of two <54>. The use of some of these designs has been simulated, and the results are summarised in Chapter 4.

iii) Hardware speed requirements are reduced as many computations are performed in frequency space parallel and not in time parallel. For example, most conventional filter designs require all of the coefficients to be used in the computation of each of the output samples. All these computations must therefore be carried out in one sampling period. In many systems this requirement has meant the use of very fast digital hardware to the exclusion of many simpler and cheaper 'pipe-line' processors. Most of the computations in the tree-structure are being carried out on different parts of the frequency domain at the same time by separate units acting in frequency space parallel.

iv) The speed requirements of each half-band filter unit are reduced by a factor of two at each inner stage of the filter, and the total data rate at each stage remains constant. Time-division-multiplexing techniques may therefore be employed to minimise the computational (ie. non-storage) hardware at each stage.

v) For particular filter characteristics, some of the tree-structure need not be implemented. For example, the filter characteristic in fig.3.2a does not require the full tree structure. A reduction in hardware may therefore be traded for a reduction in flexibility.

3.4.3 Disadvantages.

Hardware requirements could be excessive, especially in a high-resolution three-dimensional filter, because of the large amounts of storage required (field stores, line stores etc.). However, the use of techniques such as those described in the previous section and the implementation of VLSI field stores (eg. the CCD field stores now under development by Philips), can reduce the hardware requirements considerably.

3.5 Theoretical Background.

3.5.1 Introduction.

The subject of sample rate conversion in digital signal processing systems has be covered extensively by many authors <55,56,57,58,59> and therefore only a brief summary of interpolation and decimation in signal processing is given. The choice of decimation and interpolation structures is discussed, and the general form of the one-dimensional filter units (section 3.3) is shown to be a particular example of a polyphase structure. The use of these structures in integer sample rate conversion is also covered.

3.5.2 Decimation and Interpolation in Digital Signal Processing.

The digital process of reducing an input sampling rate, F1, to an output sampling rate F2, ie.,

$$F2 < F1$$

is known as decimation. The process of decimating a signal x(n) by an integer factor M is shown in fig.3.14. To avoid alias components in the output signal, y(m), the input signal should be lowpass filtered by the ideal filter characteristic :

$$H(w) = \begin{cases} 1, & |w| \le \pi/M \\ 0, & \text{otherwise.} \end{cases} \qquad \ldots.(3.2)$$

The sampling rate of the bandlimited signal, w(n), is then reduced by retaining only every Mth sample, forming the new output sequence y(m), given by

$$y(m) = \sum_{k=-\infty}^{+\infty} h(k).x(Mm-k) \qquad \ldots.(3.3)$$

The typical form of each of these signals in the frequency domain is shown in fig.3.14.

The digital process of increasing an input sampling rate, F1, to an output sampling rate F2, ie. :

$$F2 > F1$$

is known as interpolation. The process of interpolating a signal x(n) by an integer factor L is shown in fig.3.15. The sampling rate of the signal x(n) is increased by the factor L by inserting L-1 zero-valued samples between each sample of x(n), giving the signal:

$$w(m) = \begin{cases} x(m/L), & m = 0, +/-L, +/-2L, \text{ etc} \\ 0, & \text{otherwise.} \end{cases} \qquad \ldots.(3.4)$$

As shown in fig.3.15b, the spectrum of w(m) contains unwanted images of the baseband spectrum, centred at harmonics of the original sampling frequency (+/- $2\pi/L$, $4\pi/L$, ...etc.). Therefore, to recover the baseband signal, it is necessary to lowpass filter the signal w(m) with the ideal characteristic:

$$H(w) = \begin{cases} L, & |w| \leq \pi/L \\ 0, & \text{otherwise} \end{cases} \qquad ....(3.5)$$

In effect, the lowpass filter is performing the interpolation process necessary to compute the final value of the zero-valued samples in w(m).

The general case of sample rate conversion by a rational factor, M/L, is shown in fig.3.16a. The interpolation process must be carried out first, followed by the decimation process in cascade. The two lowpass filters of figs.3.14 and 3.15 would have been operating at the same sampling frequency in the cascaded structure, and therefore a more efficient implementation may be obtained by combining the two lowpass filters and using a filter with a cutoff frequency equal to the minimum of the two original cutoff frequencies (fig.3.16a).

The general process of digital sampling rate conversion is shown in fig.3.16b. It is an example of a linear time-varying system, and the system may be described by the general expression <59>:

$$y(m) = \sum_{n=-\infty}^{+\infty} g_m(n)x([mM/L] - n) \qquad \text{....(3.6)}$$

where [u] denotes the integer less than or equal to u. A number of observations can be made :

i) The system is linear as the output samples are a linear combination of the input samples.

ii) $g_m(n)$ is the response of the system at the output sample time m, to an input sample at the input sample time [mM/L].

iii) It may be shown <59> that $g_m(n)$ is periodic in m with period L ie. :

$$g_m(n) = g_{m+rL}(n), \qquad \text{....(3.7)}$$

where r = 0,+/-1,+/-2, etc

The system is therefore an example of a linear, periodically time-varying system.

iv) In the simple case, where L = M = 1, the period of $g_m(n)$ is 1 and the integer part of m - n is m - n. The equation then reduces to the familiar time-invariant digital convolution equation :

$$y(m) = \sum_{n=-\infty}^{+\infty} g(n)x(m - n) \qquad \text{....(3.8)}$$

3.5.3 Introduction to Decimation and Interpolation Structures.

The choice of implementation structure is very important when designing decimation and interpolation systems. Crochiere and Rabiner <55> have given an summary of various structures in their paper. This section concentrates on the technique of transposition, which is of importance to the new filter design technique. The important technique of commutation and examples of Direct Form and Polyphase structures are then discussed.

The principles of duality and transposition are important concepts in digital signal processing. Claasen et al. <60> have shown that decimation and interpolation are dual processes and that decimators may be obtained from interpolators, and vice versa, using transposition techniques. This important result is used later when the design of the one-dimensional half-band filter networks is considered.

To form the transpose of a network structure, the original signal flowgraph is reversed and each element is replaced by its transpose. Linear time-invariant elements, such as gains and delays, remain unchanged. Furthermore, in the case of linear time-invariant systems, the input-to-output response of the original system and its dual are identical <60>. For example, the well-known signal flowgraph for a direct-form FIR filter is shown as part of fig.3.17a, The input-to-output responses of this structure and its transpose are identical. Time-varying systems, such as decimation and interpolation, do not possess this property.

3.5.4 Direct-form FIR Structures.

A simple direct form FIR structure for integer decimation may be derived easily from the basic form of a decimator shown in fig.3.14, and the basic direct form FIR filter structure. Replacing the lowpass filter h(n) with the filter flowgraph leads to the structure in fig.3.17a. A direct implementation of this sample rate conversion structure would be very inefficient for a number of reasons :

i) The lowpass filter is operating at the higher sampling rate.

ii) Only one in every M samples is required at the output of the filter.

However, the branch operations of sample rate reduction and gain may be commuted, leading to the more efficient structure shown in fig.3.17b. Coefficient multiplications and additions now occur at the lower sampling rate F/M and therefore the total computation rate has been reduced by a factor of M.

The importance of linear phase has been discussed in section 1.1.4, and in the case of FIR filters, the impulse response must be symmetrical for linear phase. This symmetry may be exploited by combining multiplications with the same coefficient. Excluding the central coefficient (dc term), the number of multiplications is therefore reduced by a factor of two. This well-known advantage of symmetrical filters is exploited further when implementations for the tree-structure units are suggested in section 3.6.

An efficient structure for integer interpolation may be derived in a similar way. Using the duality property, the interpolation structure could be obtained directly from the decimation structure by direct transposition. Classen et al. <60> also made two further interesting observations :

i) The complexity of both systems, when expressed in terms of the number of multiplications or number of stages etc., is the same.

ii) The sampling rate at which the multiplications take place will not change.

Therefore the efficiency of the transposed structure is directly related to that of the original structure.

3.5.5 FIR Polyphase Structures.

One particular implementation of the one-dimensional filter units is an example of an FIR Polyphase (or N path) network (section 3.6.1).

Bellanger et al. <61> have shown that the FIR filter of fig.3.17 could be implemented using a polyphase structure. Fig.3.18 shows a polyphase filter network of order N. The system transfer function is given by:

$$H(Z) = \sum_{K=0}^{R-1} Z^{-K} . P_K(Z^R) \qquad \text{....(3.9)}$$

where R is the decimation factor (M) or the interpolation factor (L). The network performs the filtering operation by shifting the phase of the input signal in such a way as to allow unwanted frequency components to be cancelled at the output of the summation device. Using the efficient structures developed for decimation and interpolation, the equivalent structures for polyphase networks may be derived easily (fig.3.19). Again, both the structures are efficient as the polyphase filters are functions of Z^R and are therefore operating at the lower sampling rate. The impulse responses of the individual polyphase filters are in fact decimated versions of the original lowpass filter h(n). For example, the ideal lowpass filter for decimation is shown in fig.3.20a, and the corresponding characteristic for the polyphase filters is shown in fig.3.20b. The polyphase filters are in fact all-pass sections, and in general, each polyphase filter may exhibit a different phase shift. Therefore, in practice, each filter has a different delay and these delays are compensated for by the delays occurring at the higher sampling rates. Constantinides and Valenzuela have investigated the use of these allpass decimation and interpolaton networks in transmultiplexing <52,62>.

The sample rate expanders/compressors and sample delays in fig.3.19 may be replaced by commutators (figs.3.21, 3.22). In fact, the allpass network implementations described later, are examples of these commutator models.

3.6 FIR Polyphase Structures Utilising Half-band Filters.

A fuller analysis of the tree-structure filters introduced in the last section is given in the following sections. A pictoral as well as mathematical analysis is given in an attempt to aid the understanding of the fundamental operation of the filter.

Each stage in the efficient tree-structure filter of fig.3.13 represents a decimation or interpolation process by a factor of two and efficient structures for sample rate conversion by factors of two are also described. Finally, a specific practical implementation for the filter units, using FIR half-band filters, is analysed. Further examples of these units are given in chapter 4.

3.6.1 Decimation and interpolation by factors of two.

The tree-structure implementation of the new filter structure, described in section 3.3, utilised one-dimensional decimation and interpolation units in a tree pattern. The general forms for the decimator and interpolator units are shown, together with the appropriate frequency domain characteristics, in fig.3.23. Each unit is used effectively to either increase or decrease the input sampling rate by a factor of two.

The polyphase commutator models for decimation and interpolation by factors of two (fig.3.24) may be derived easily from the general models (figs.3.21, 3.22) by placing M=L=2. From the general transfer function (equn.3.9), the transfer function of the individual units is given by:

$$H(z) = P_0(z^2) + z^{-1}.P_1(z^2) \qquad \ldots.(3.10)$$

In the decimator units (fig.3.23a), both lowpass and highpass outputs are required at one-half the input sampling rate. The lowpass output may be obtained by a direct implementation of the polyphase network (fig.3.25a). The transfer function of the structure may be derived by setting $z = e^{j\omega}$ in equn.3.10 :

$$A(w) = \left| e^{j\phi_0(\omega)} + e^{j\phi_1(\omega)} \right|$$

$$= \left| e^{j\phi_1(\omega)} \left(1 + e^{j\{\phi_0(\omega)-\phi_1(\omega)\}} \right) \right|$$

$$= 2.\left| \cos\{(\phi_0(w) - \phi_1(w))/2\} \right| \qquad \text{....(3.11)}$$

The transfer function is therefore lowpass, as required.

It is well known that a highpass filter may be transposed from a lowpass filter by changing the sign of alternate samples (or coefficients). The highpass output is therefore obtained by subtracting the samples at the outputs of the two polyphase filters (fig.3.25b). From equn.3.10, the corresponding transfer function is given by :

$$H(w) = \left| e^{j\phi_0(\omega)} - e^{j\phi_1(\omega)} \right|$$

$$= \left| e^{j\phi_1(\omega)} \left(1 - e^{j\{\phi_0(\omega)-\phi_1(\omega)\}} \right) \right|$$

$$= 2.\left| \sin\{(\phi_0(w) - \phi_1(w))/2\} \right| \qquad \text{....(3.12)}$$

The transfer function is therefore highpass, as required.

A polyphase structure to implement the complete decimator unit of fig.3.23a is shown in fig.3.26a. The structure is basically a combination of the lowpass and highpass decimator structures (fig.3.25). However, the two phase shifters are combined to give the following differential phase shift (fig.3.27):

$$e^{j\phi} = e^{j\phi_0} - e^{j\phi_1} \qquad \text{....(3.13)}$$

Applying the principle of duality, the interpolator unit of fig.3.23b may be obtained from the decimator unit by transposition (fig.3.26b).

The half-band filter is in fact an efficient tool for sample rate conversion by factors of two <57>. Because of the importance of linear phase (section 1.1.4), only symmetrical FIR half-band filters will be considered at this stage. A symmetrical FIR half-band filter has linear phase and a non-causal impulse response given by :

$$h(n) = \{\sin(n\pi/2\}/n \qquad \text{....(3.14)}$$

The impulse response is symmetrical about the time origin and is zero at non-zero integer multiples of the twice the sampling period. Therefore every non-zero even index coefficient is null, resulting in a possible reduction in the number of computations over other symmetrical FIR designs of :

$$\text{computational reduction} = (N - 1)/4 \qquad ....(3.15)$$

Every output sample of the half-band filter is obtained from the input samples by the summation :

$$y(n) = x(n) + \sum_{i=1}^{M} h(2i-1).\{x(n+2i-1) + x(n-2i+1) \qquad ...(3.16)$$

where $y(n)$: output sample

$x(n)$: input sample

$M = (N+1)/4$ M,N integer

(This ensures that the first and last coefficients are non-zero)

N : filter length.

The frequency response is given by :

$$H(w) = h(0) + 2\sum_{n=1}^{M} h(2n-1).\cos\{(2n-1)w/T) \qquad ....(3.17)$$

where $1/T$ = sampling frequency.

Fig.3.28a shows the response of a typical half-band filter. Passband and stopband ripple are equal and the response is symmetrical about its 6dB point ($f_S/4$) ie :

$$| H(w) | = 1 - | H(\pi - w) | \qquad \text{....(3.18)}$$

Bellanger et al. <57> have taken advantage of the zero coefficients in the impulse response to propose half-band filter structures for decimation and interpolation by two.

The particular polyphase structures derived for the tree-structure decimator-interpolator units (fig.3.26) may also be implemented efficiently using a suitable half-band filter structure. One of these structures is derived in the following section.

3.6.2 Half-band filter implementations for the tree-structure units.

Using the decimation and interpolation structures derived in the previous sections, a simple and efficient polyphase structure has been designed which utilises FIR half-band filter designs. Fig.3.28a shows the frequency response of a typical half-band filter. If the central coefficient (or dc component) of equn.3.17 (h(0)) is removed, and the amplitude scaled by a factor of two, then the frequency response shown in fig.3.28b is obtained. The amplitude response, given by:

$$A(w) = 4.\left|\sum_{n=1}^{M} h(2n-1).\cos((2n-1)w/T)\right| \qquad \text{....(3.19)}$$

is approximately allpass (see results in Chapter 4), and the phase changes by π at $f_S/4$. This filter forms the phase-shifter ($e^{j\phi}$) in the decimation and interpolation units. The second branch of the units contains an allpass unit of gain h(0) (usually normalised to 1) to obtain the correct amplitude response at the output of each unit. In practice, the phase-shifter will have a delay given by:

$$(N+1)/2 = 2M \qquad ....(3.20)$$

where N is the length of the filter. The second branch therefore requires a delay of:

$$(N-1)/2 = 2M-1 \qquad ....(3.21)$$

to ensure coincidence of the samples in each branch. An example for N=19 is given in fig.3.29. The interpolator unit would utilise the same half-band filter, but in the configuration of fig.3.26b.

To aid the understanding of the half-band filter structure operation, the spectra for the general decimation and interpolation units are shown in fig.3.30. The result of cascading a decimator and an interpolator unit should be the identity function. Fig.3.30 shows that the interpolator unit should have a gain of 1/2 in this case. This may be achieved easily by scaling the coefficients of the half-band filter.

The advantages of this implementation have been discussed in section 3.4. Examples of various designs are given in Chapter 4.

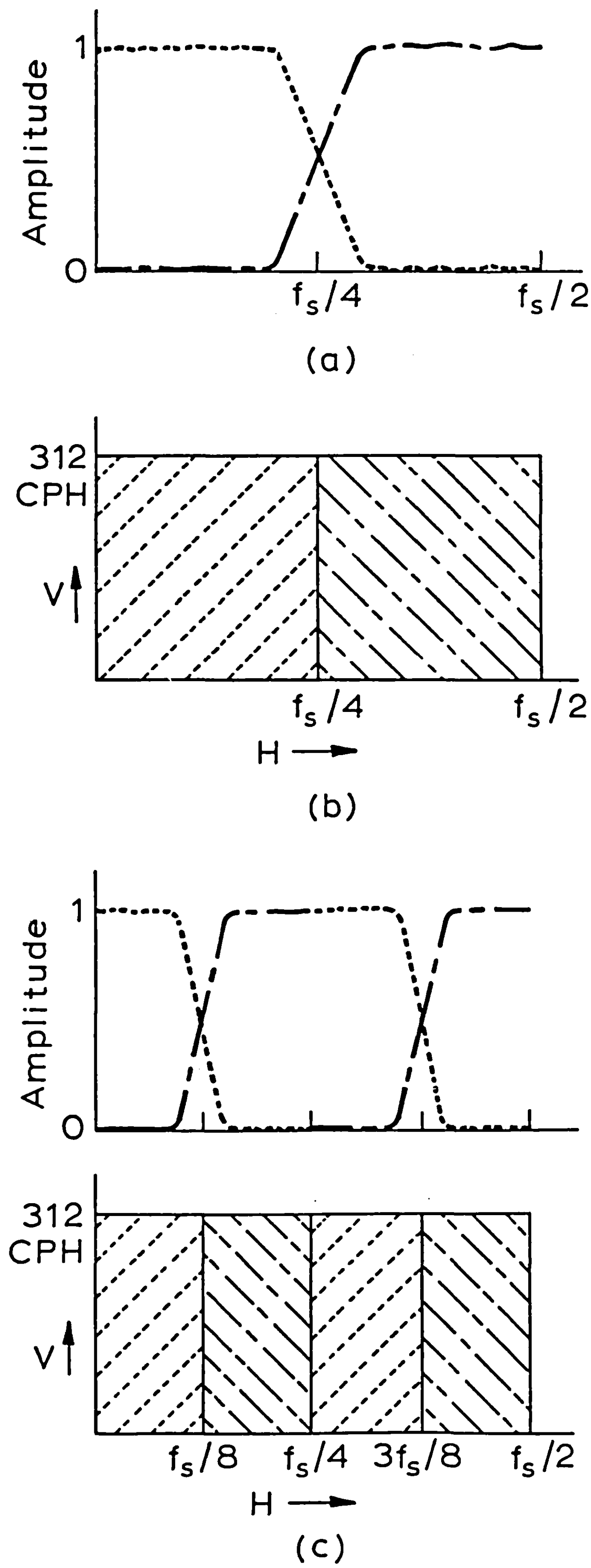
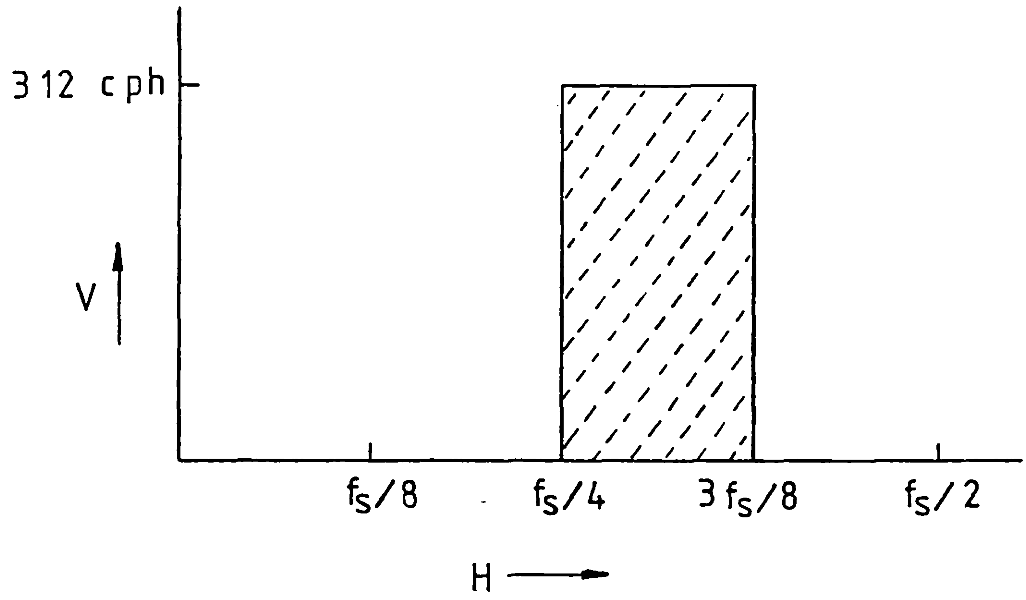
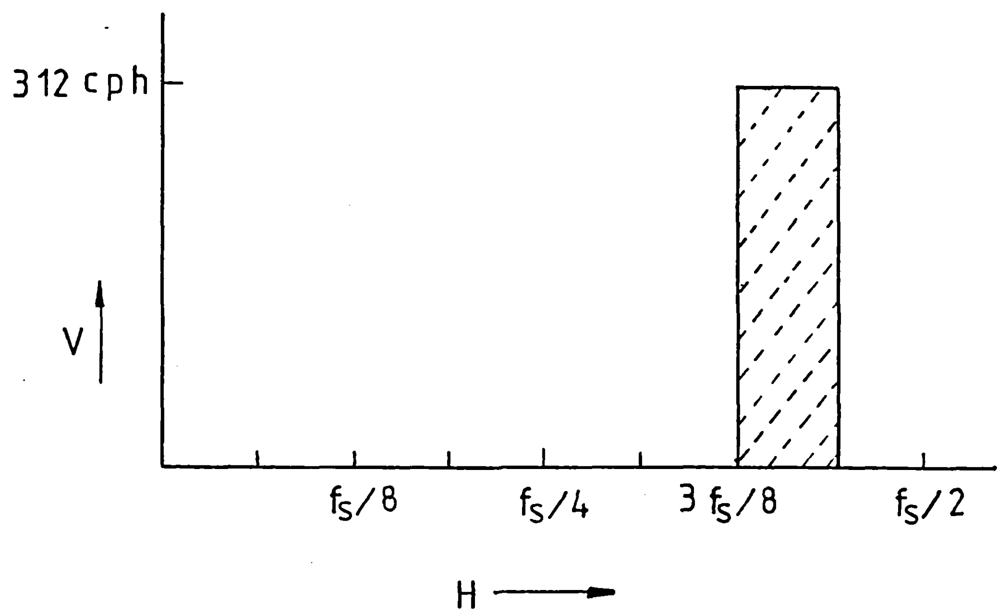


Fig. 3.1 Horizontal half-band filters.



(a)



(b)

Fig. 3.2 Bandpass filters.

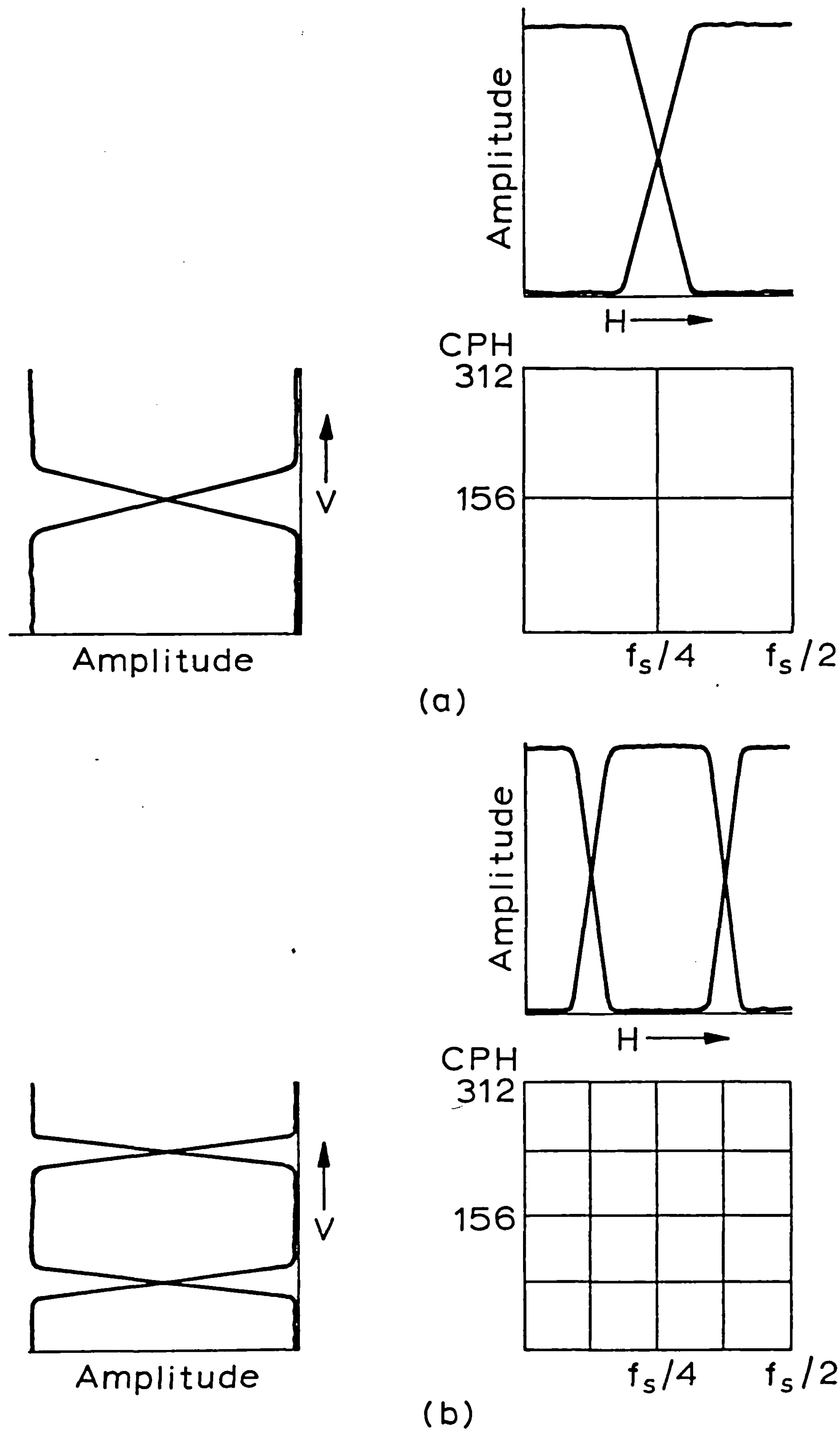


Fig. 3.3. Horizontal & vertical half-band filters.

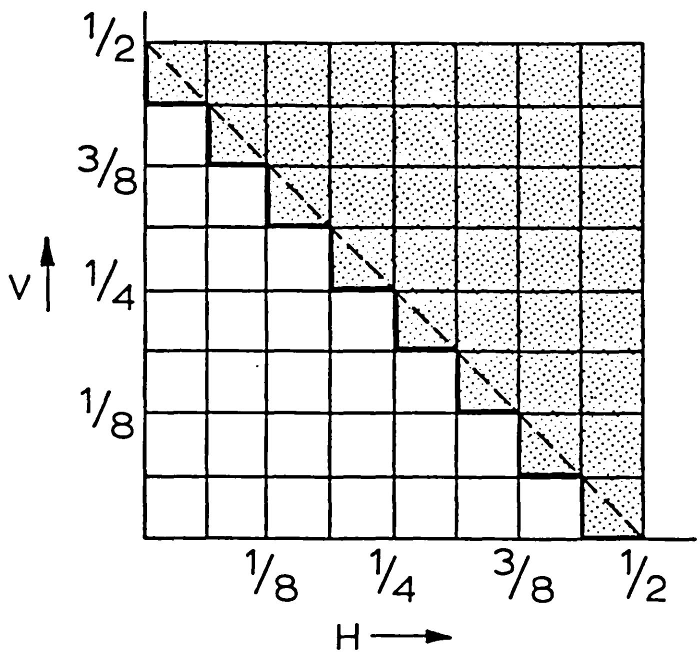


Fig. 3.4 Approximation to field quincunx response.

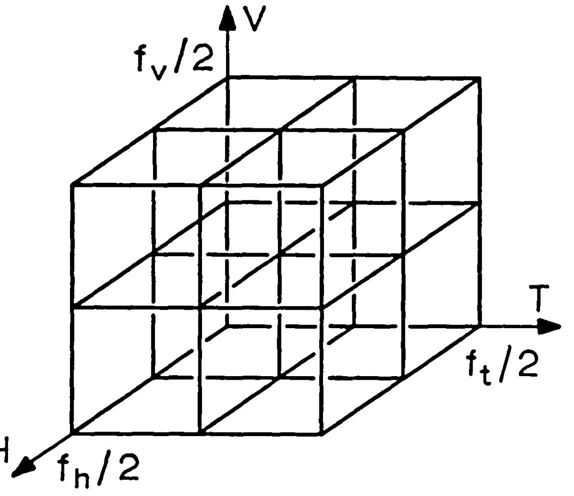


Fig.3.5 Horizontal, vertical & temporal half-band filters.

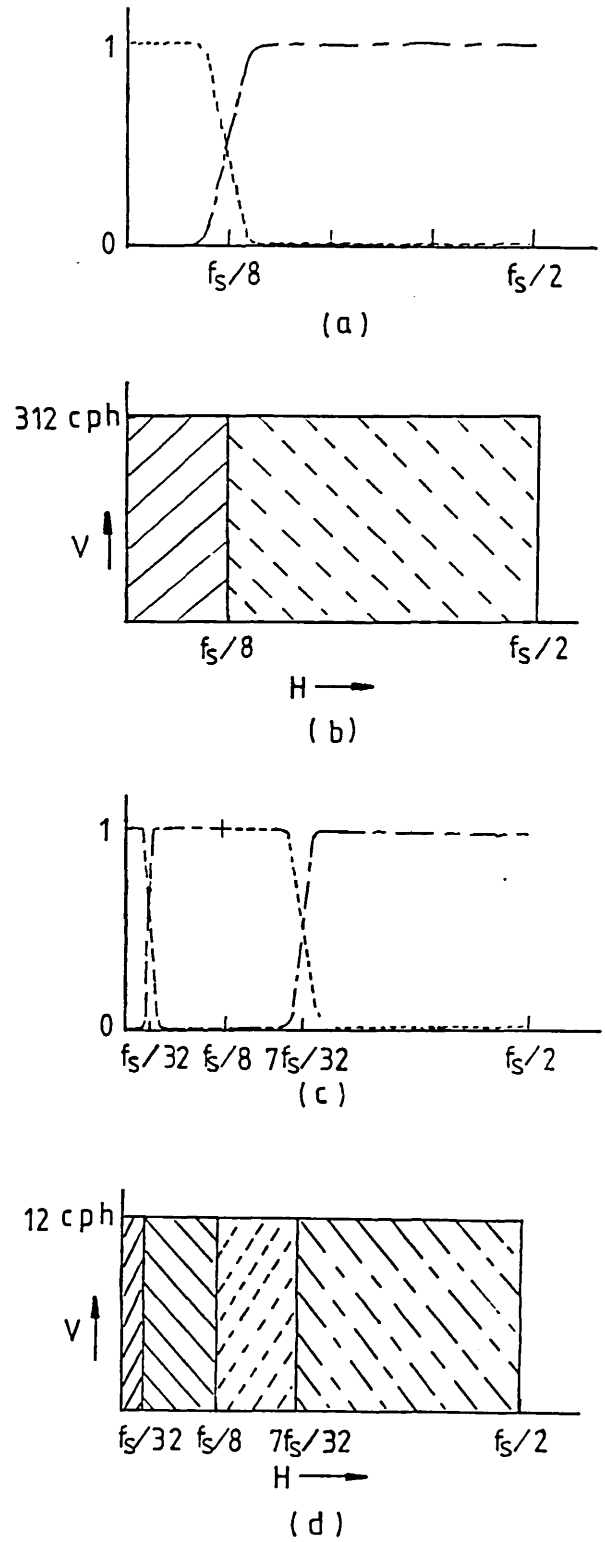
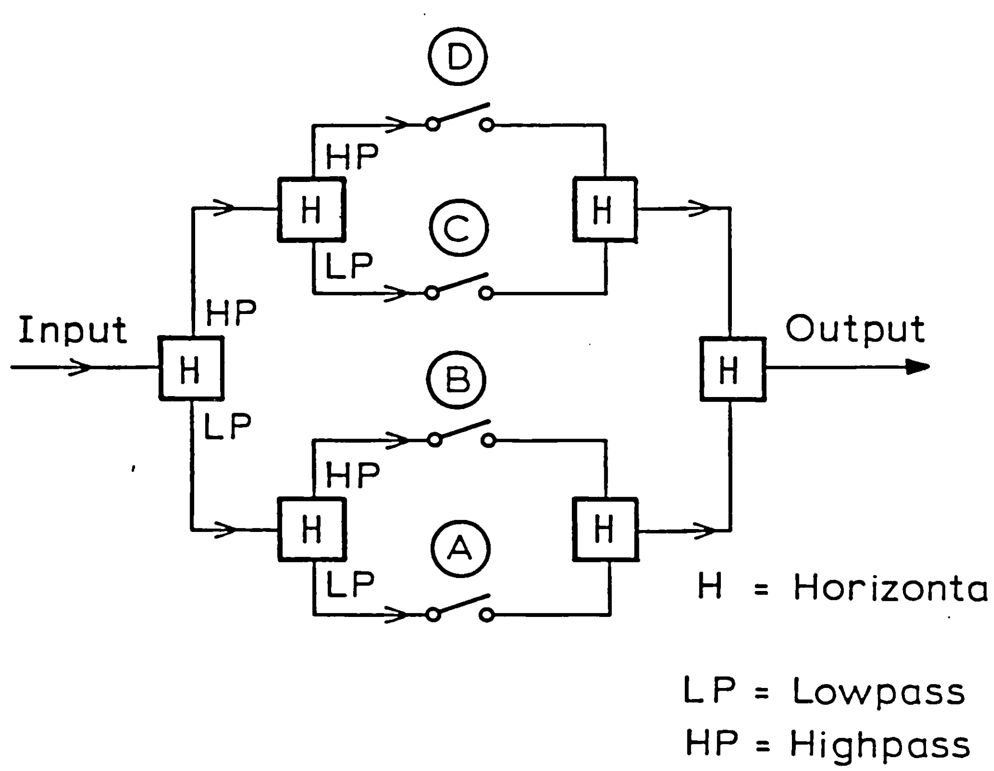
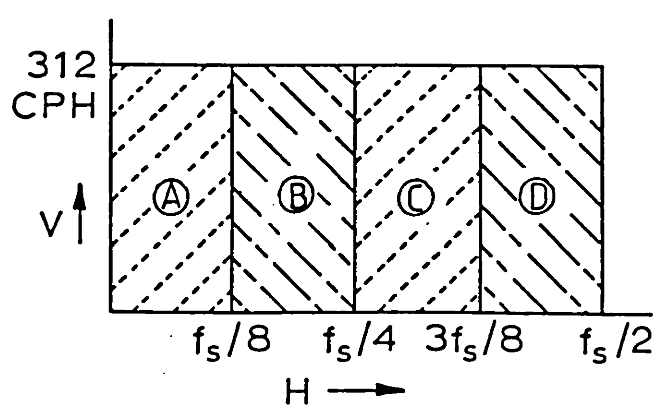


Fig. 3.6 Horizontal asymmetric filters.



(a)



(b)

Fig. 3.7 1-D filter structure.

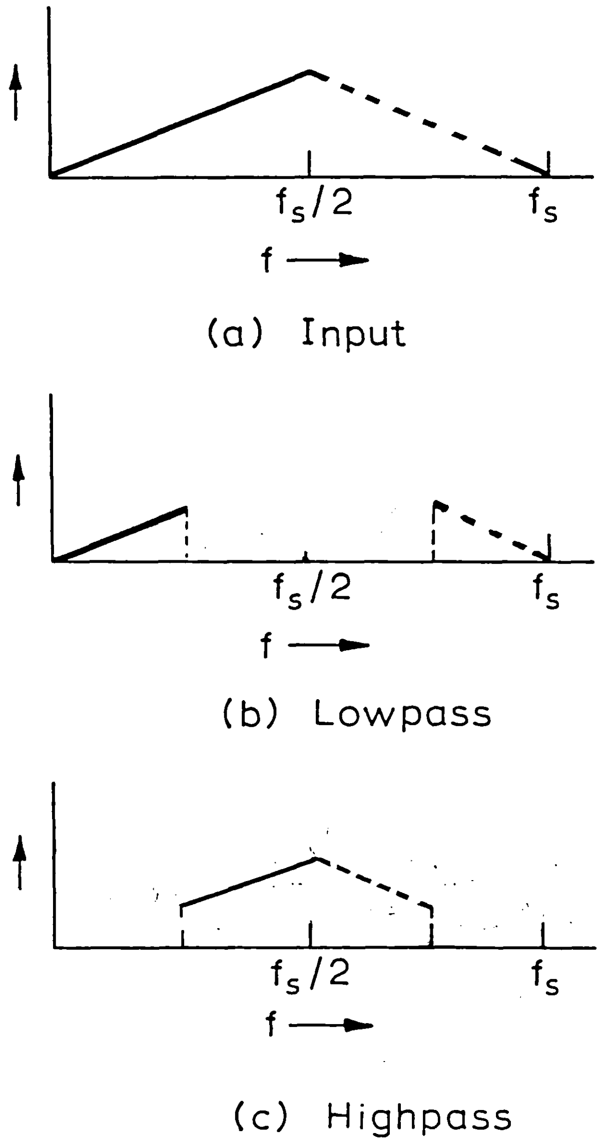
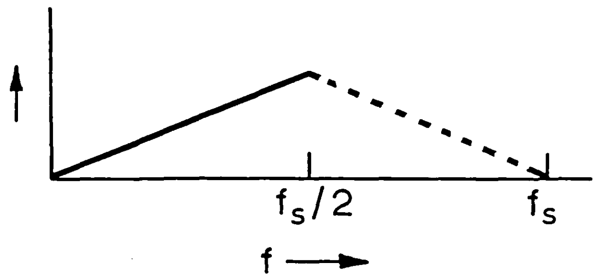
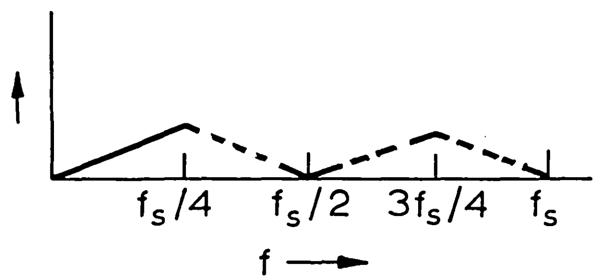


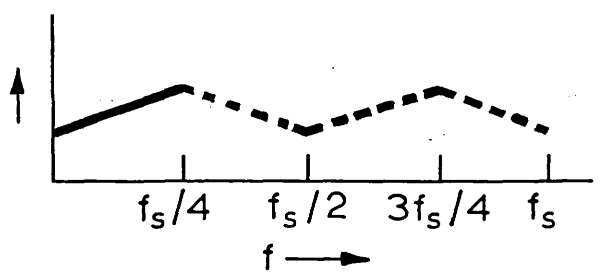
Fig. 3.8. Input & output spectra for decimation units (full rate).



(a) Input



(b) Lowpass



(c) Highpass

Fig. 3.9. Input & output spectra for decimation units (half rate).

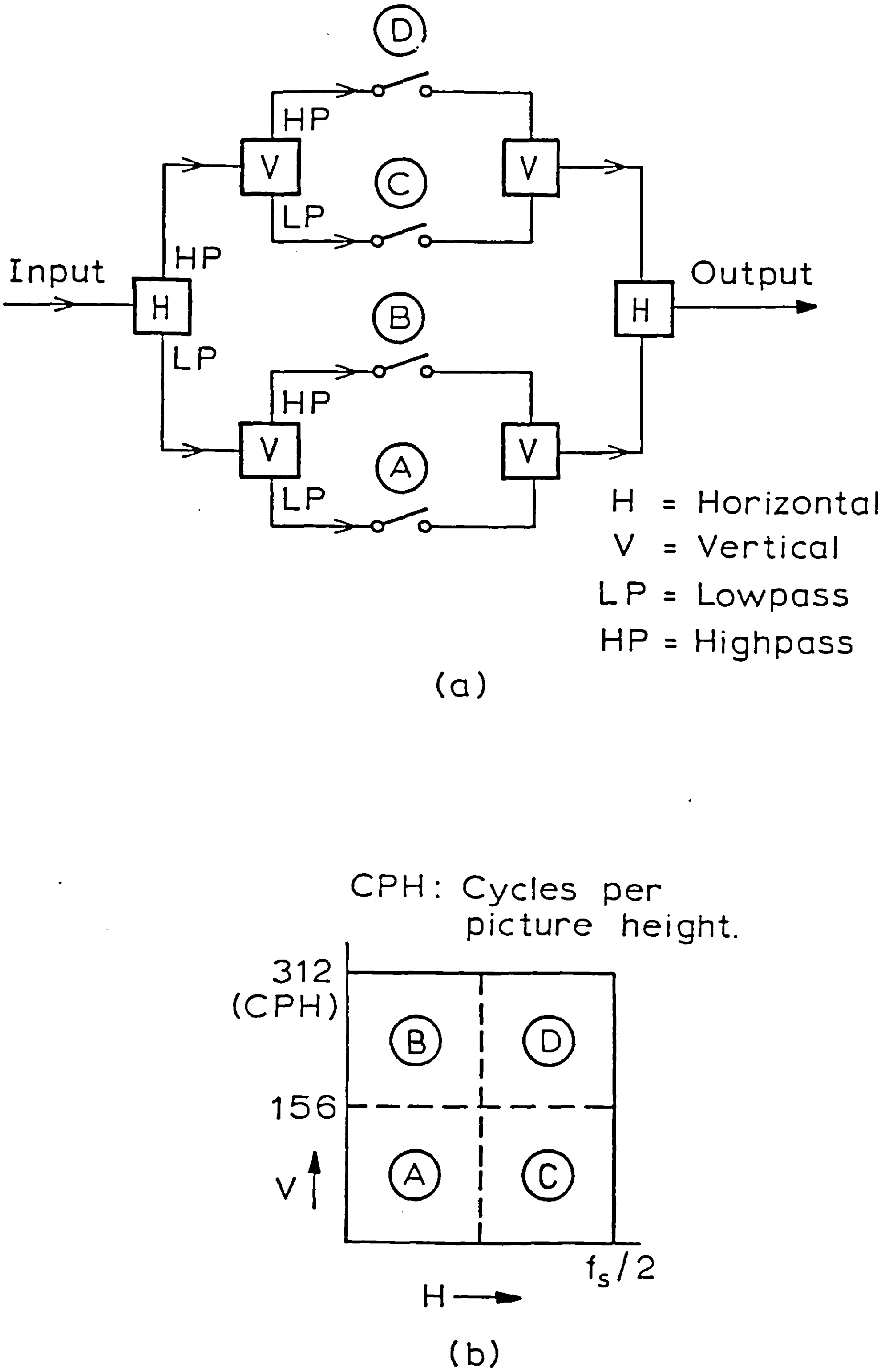


Fig. 3.10 2-D filter structure.

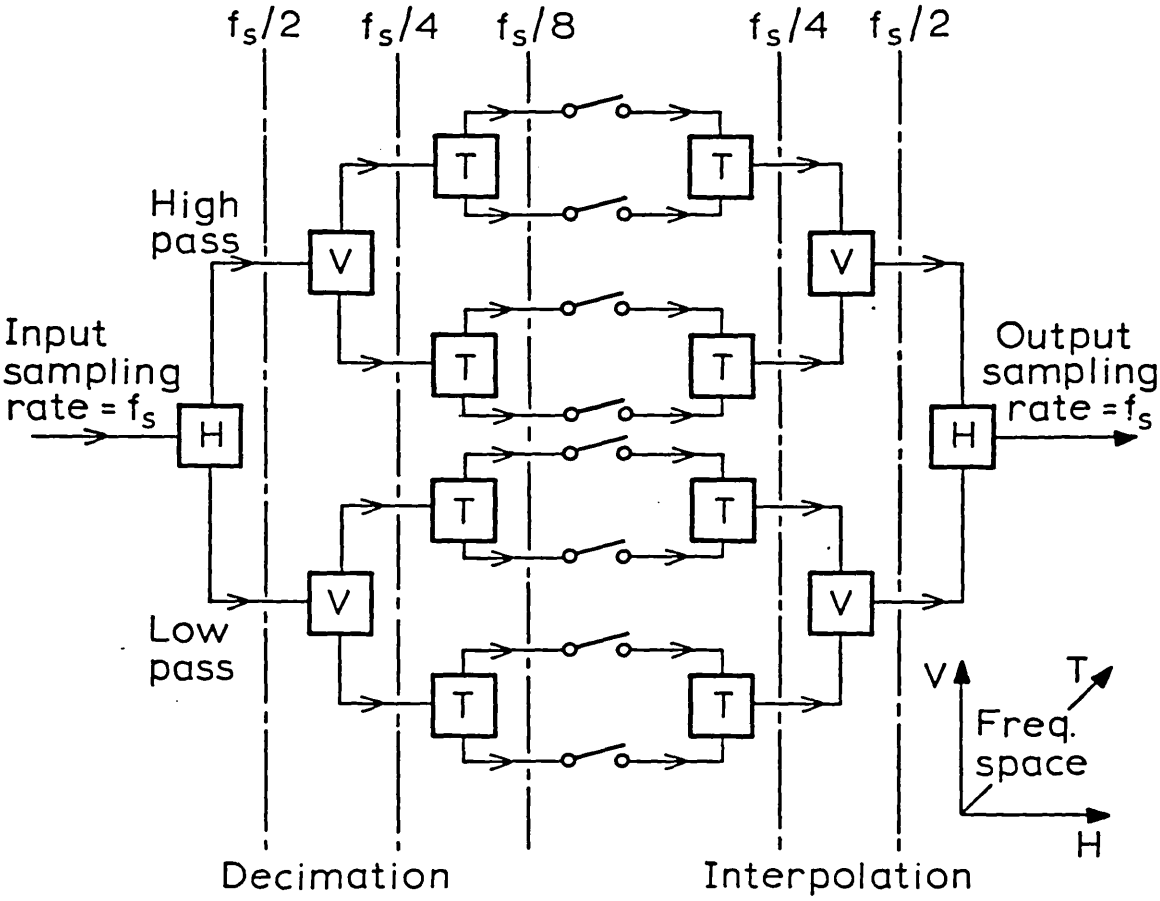
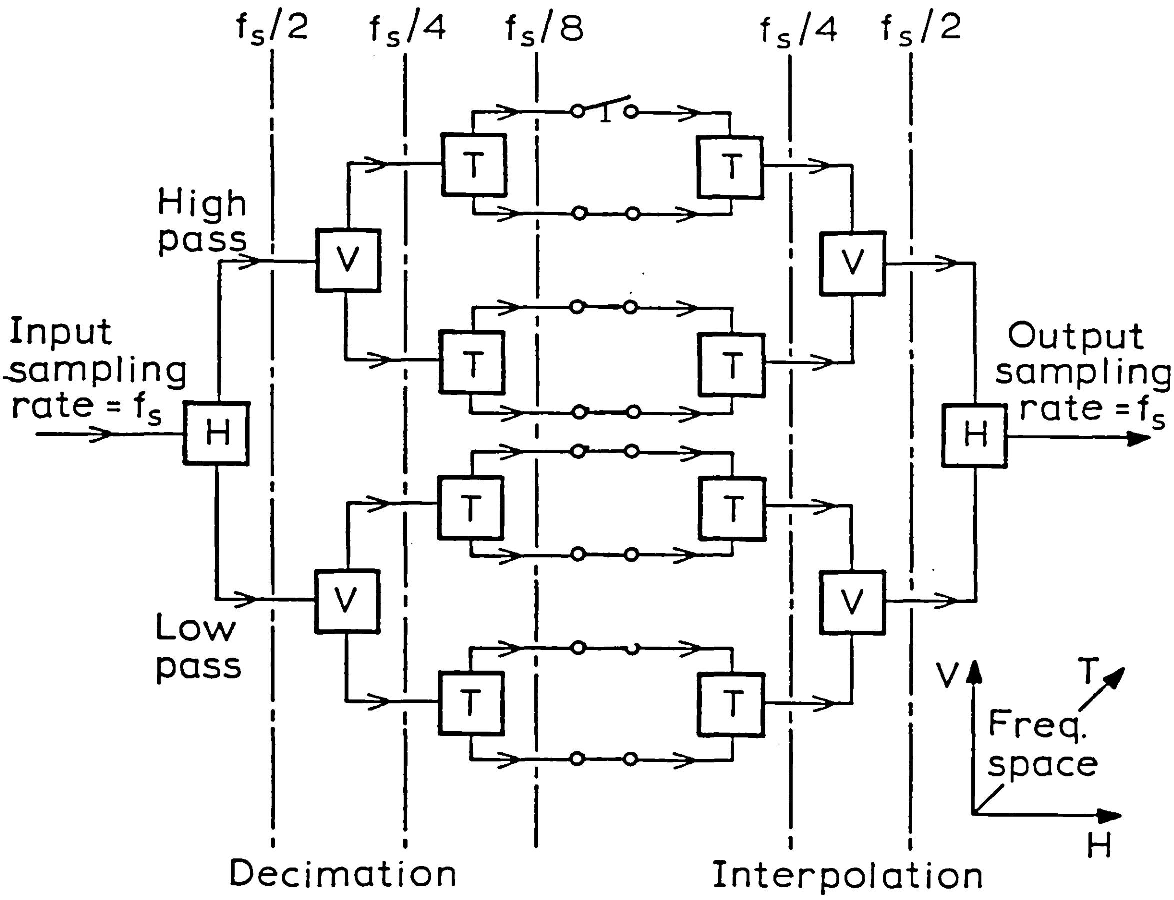
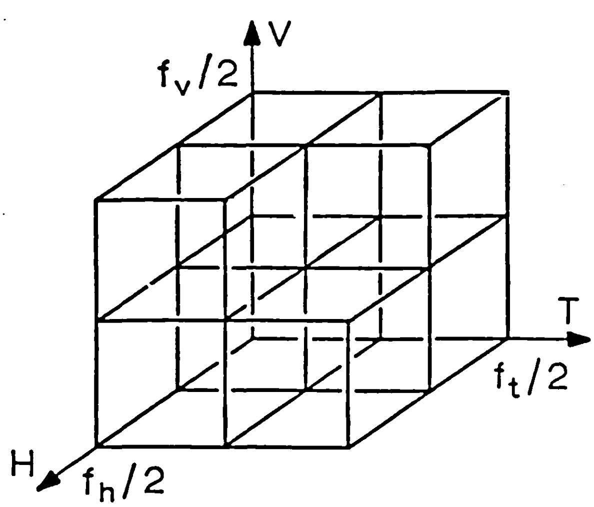


Fig.3.11 3-D filter structure.



(a)



(b)

Fig.3.12 3-D filter structure for field quincunx down sampling.

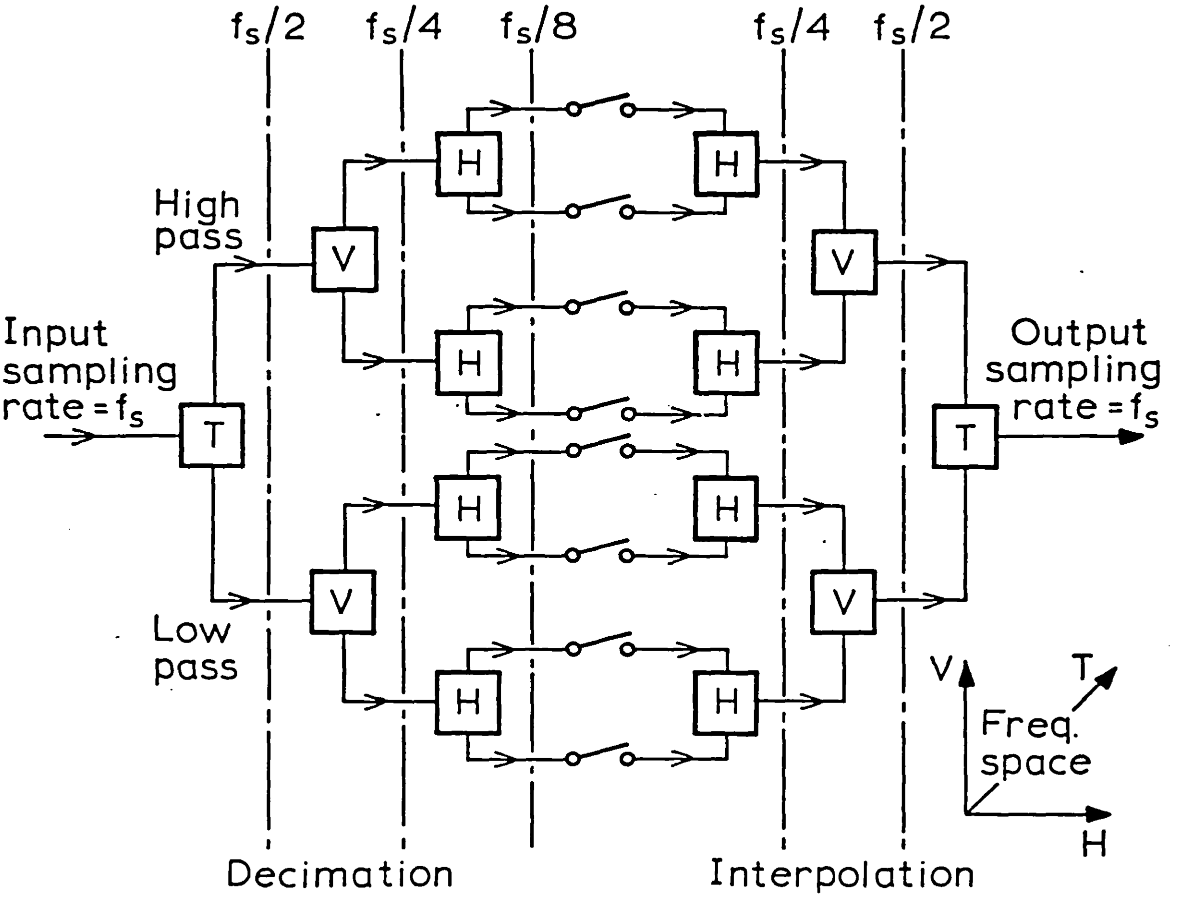


Fig. 3.13 Efficient 3-D filter structure.

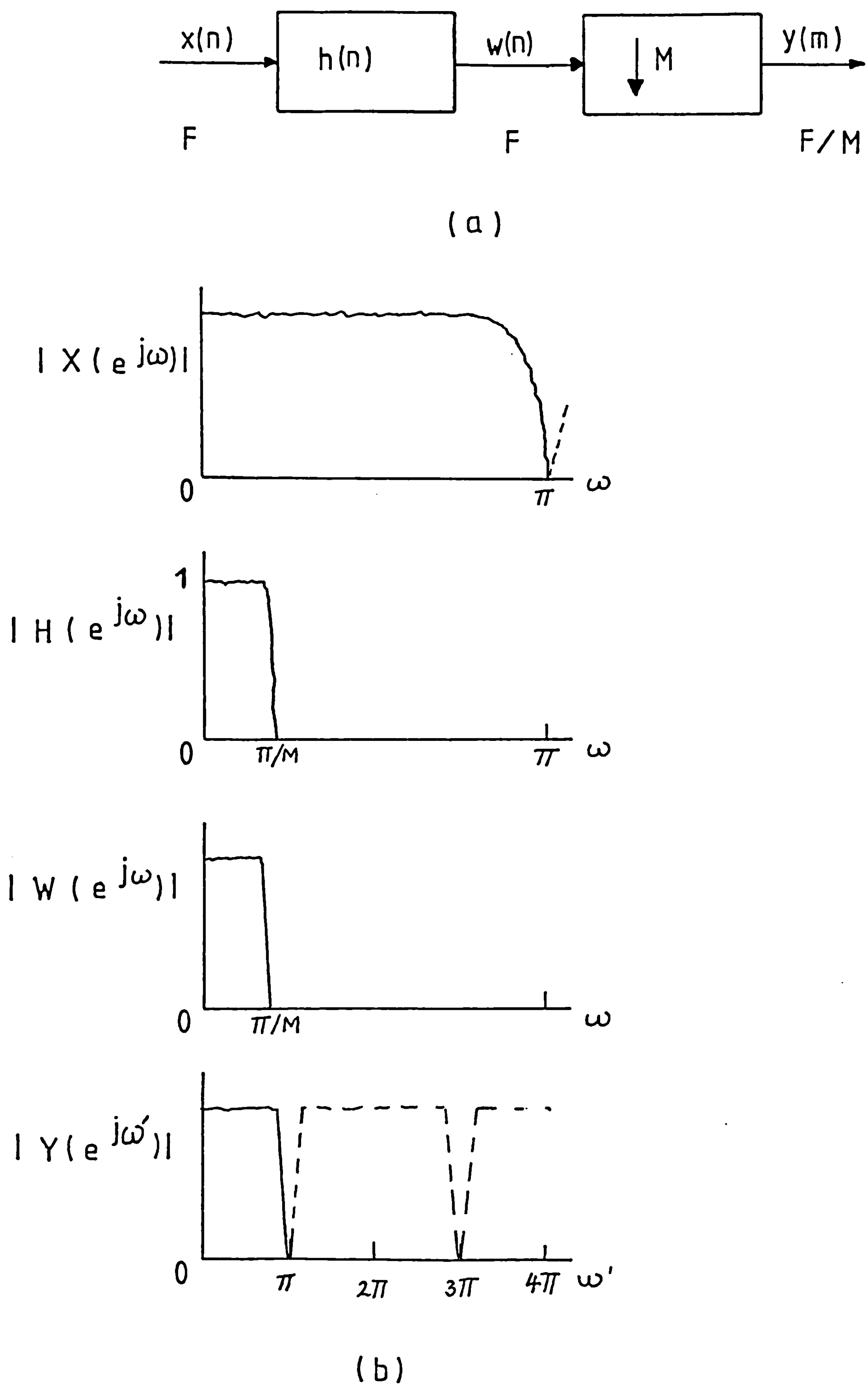


Fig. 3.14 Decimation by M.

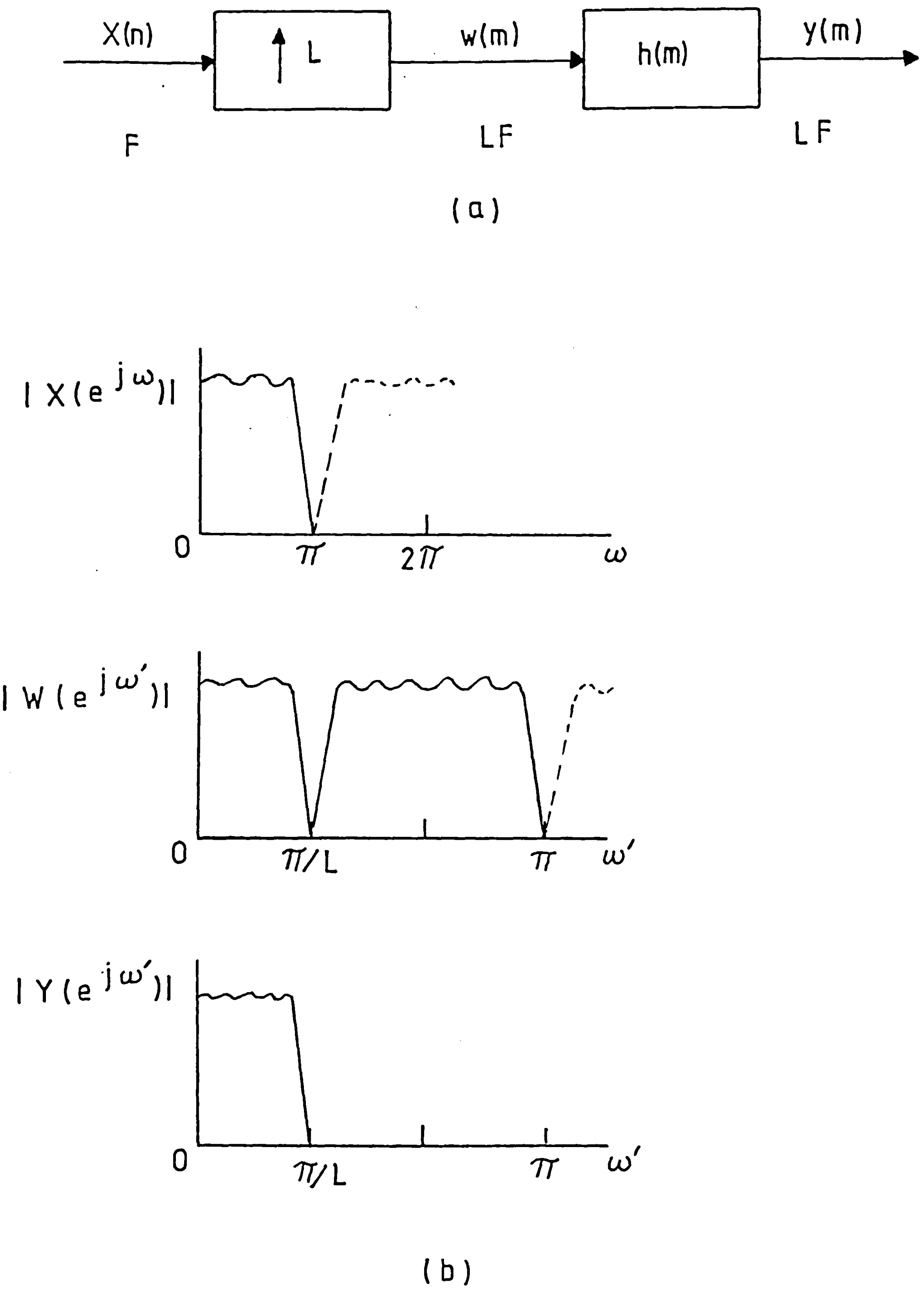
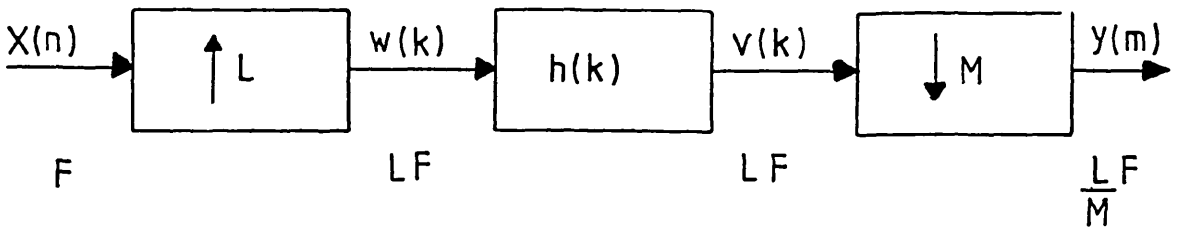
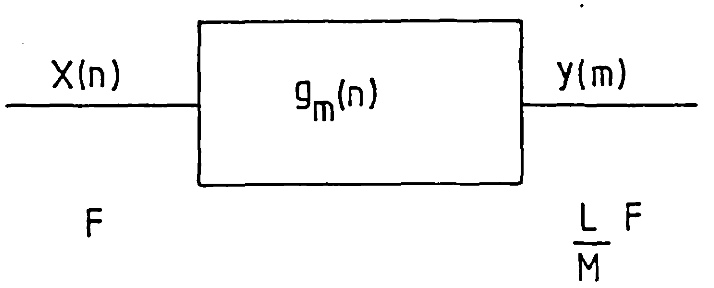


Fig. 3.15 Interpolation by L.

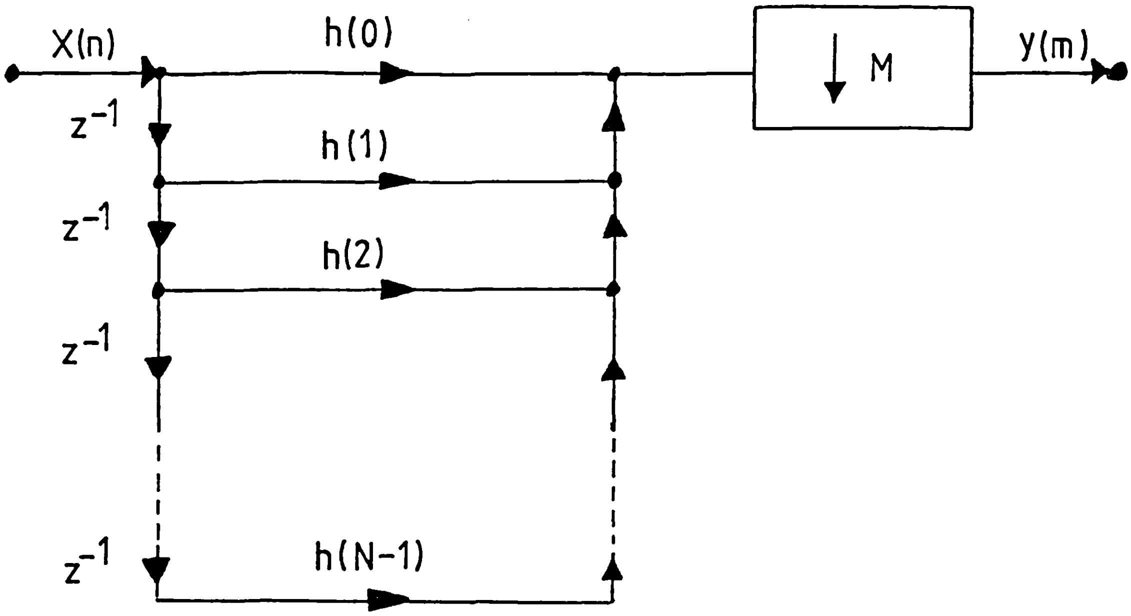


(a)

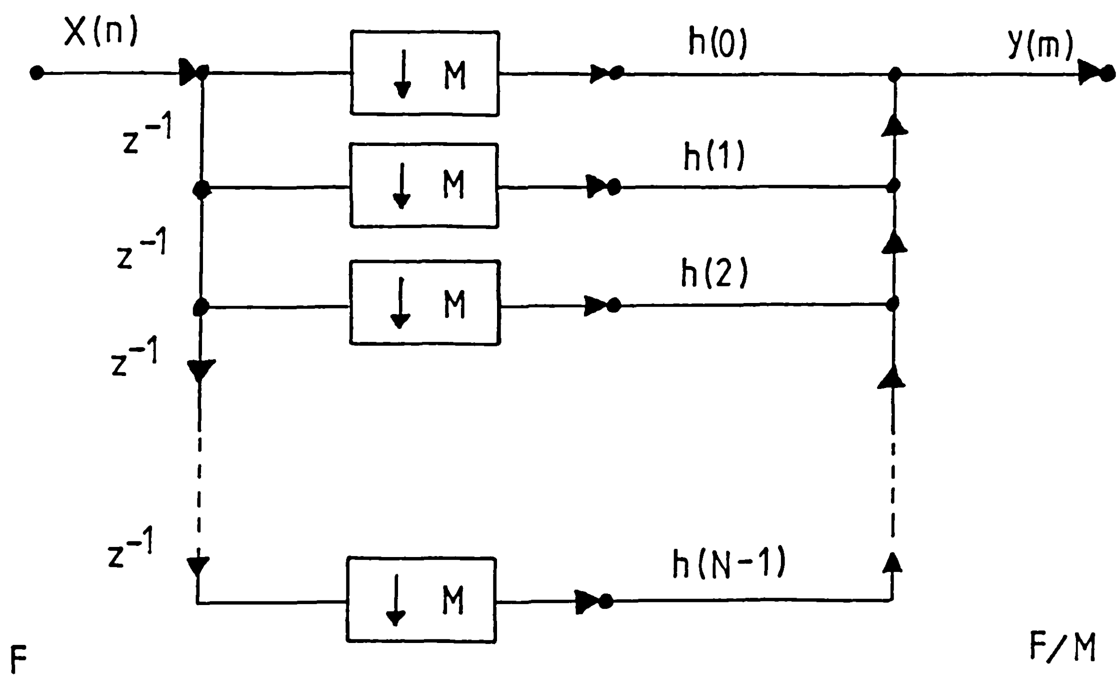


(b)

Fig. 3.16 Rational sampling rate conversion.



(a)



(b)

Fig. 3.17 Direct form structure for decimation by M.

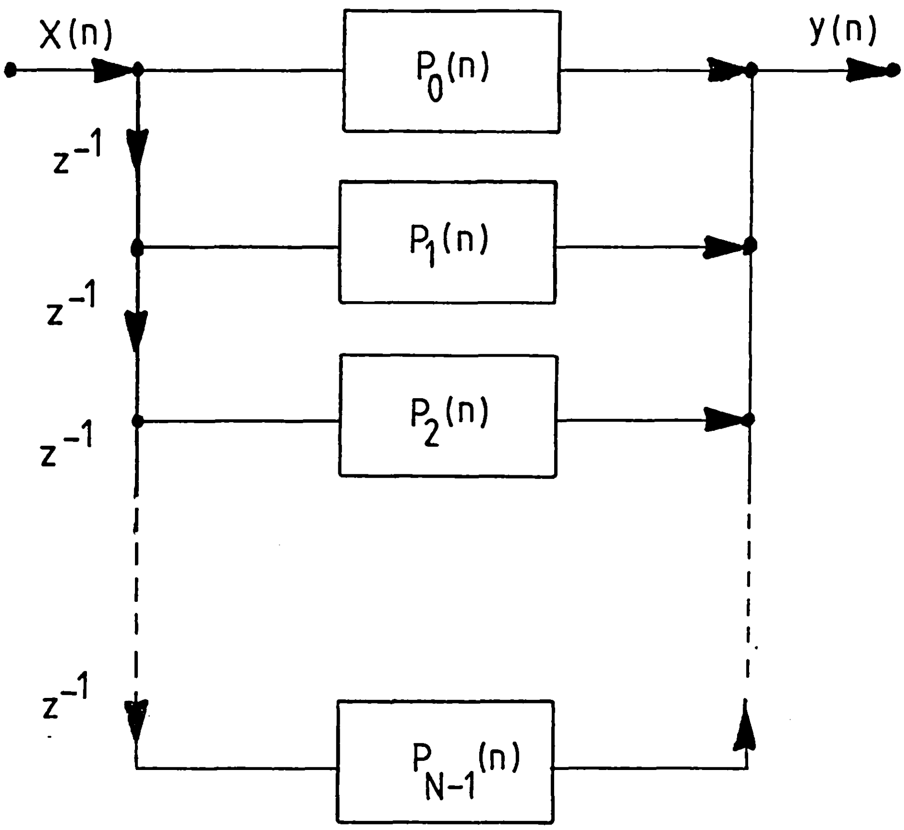
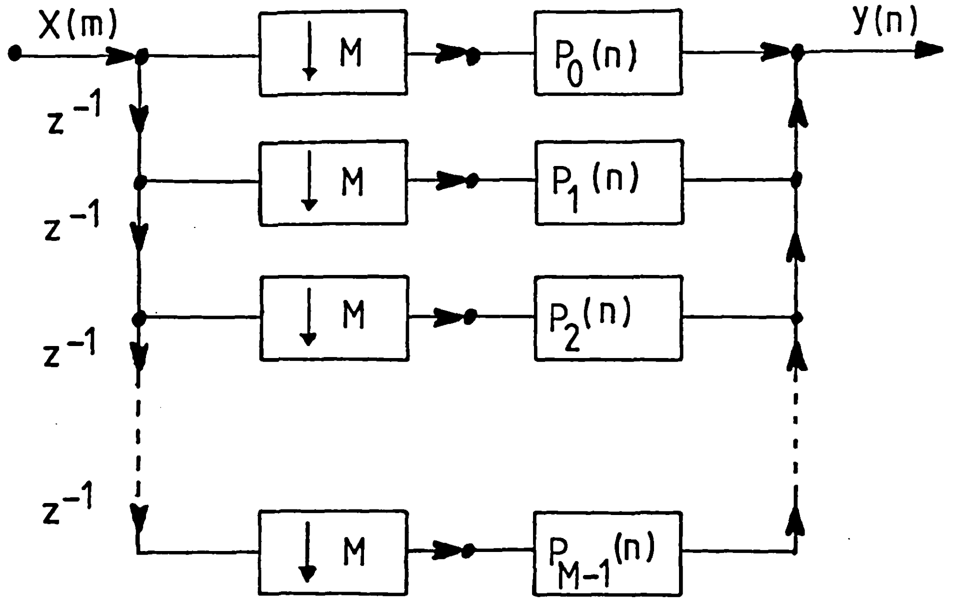
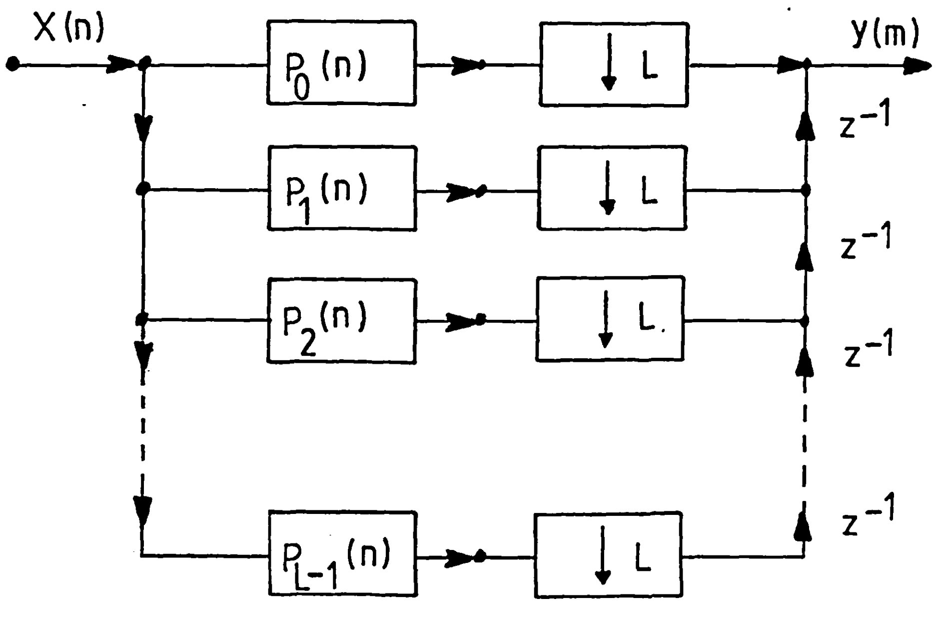


Fig.3.18. General polyphase filter structure.

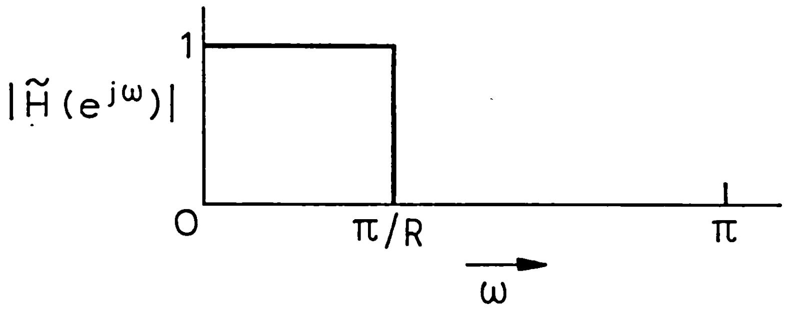


(a) Decimation.

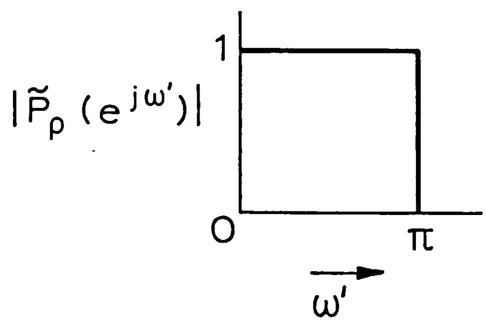


(b) Interpolation.

Fig.3.19 Polyphase networks.



(a)



(b)

Fig.3.20 Ideal frequency response of the polyphase networks.

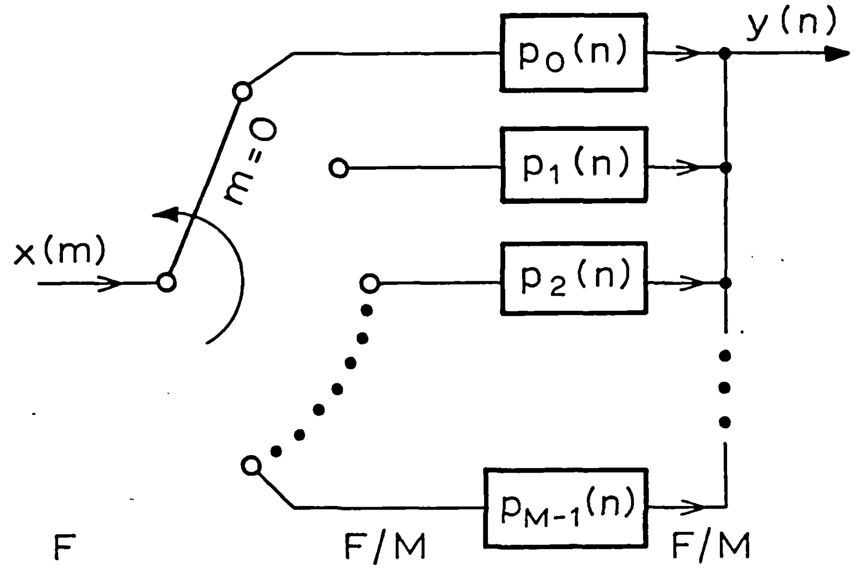


Fig. 3.21 Commutator model for the M to 1 polyphase decimator.

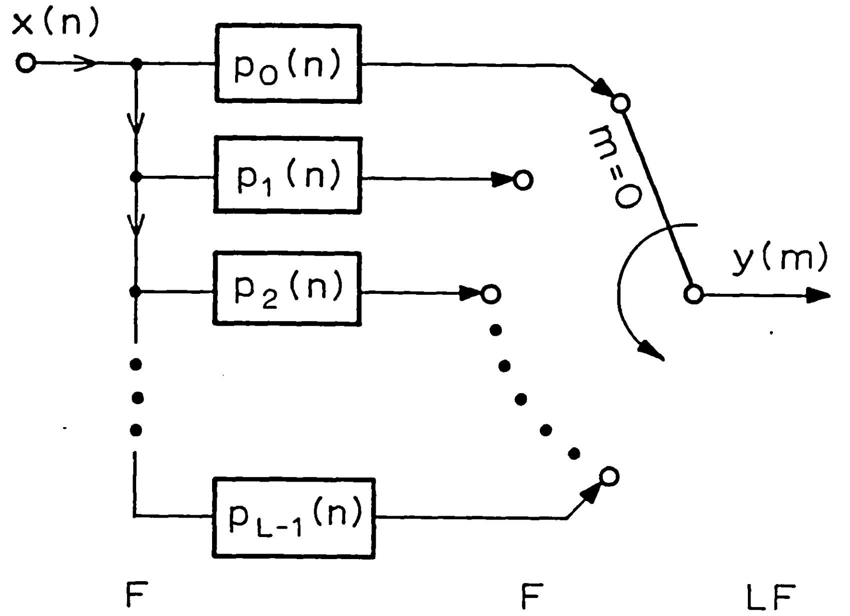
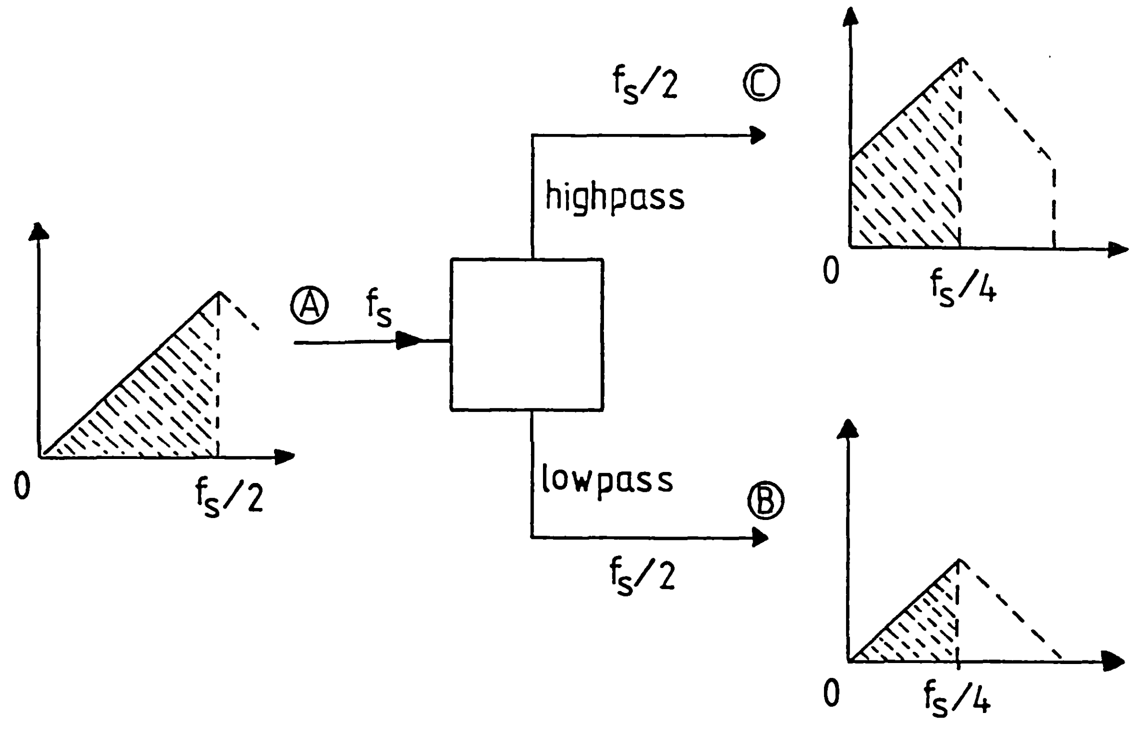
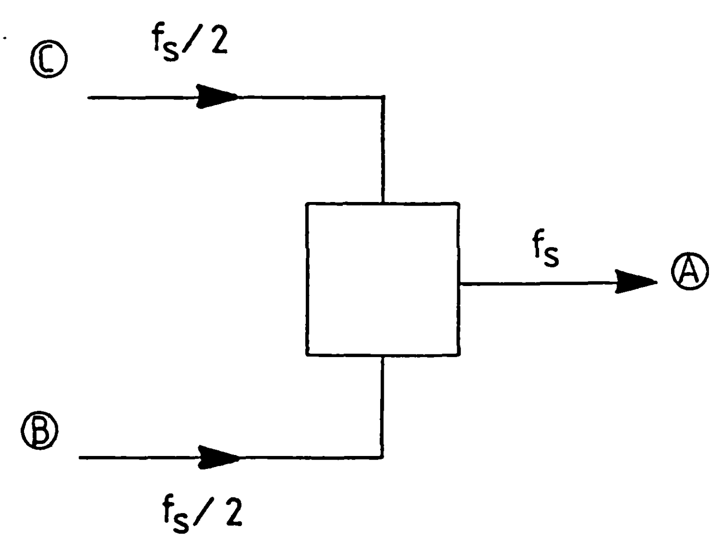


Fig.3.22 Commutator model for the 1 to L polyphase interpolator.

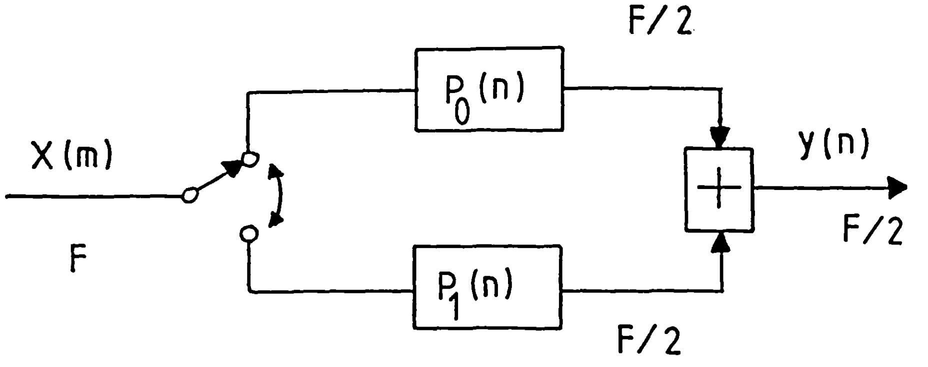


(a) Decimation

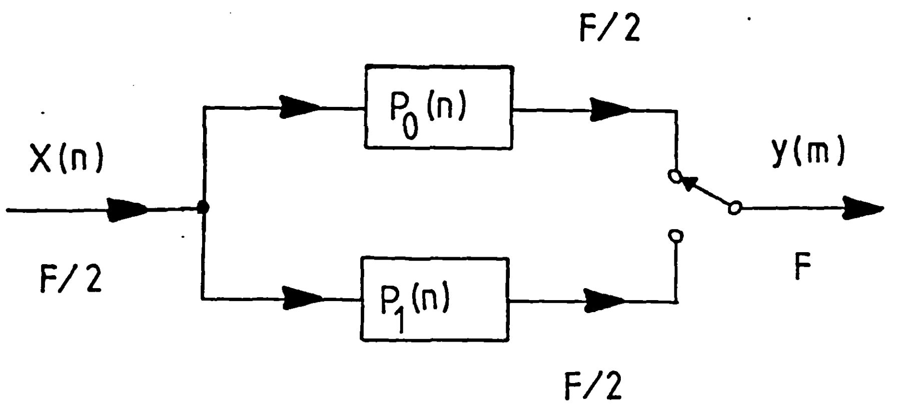


(b) Interpolation

Fig.3.23 Tree structure filter units.

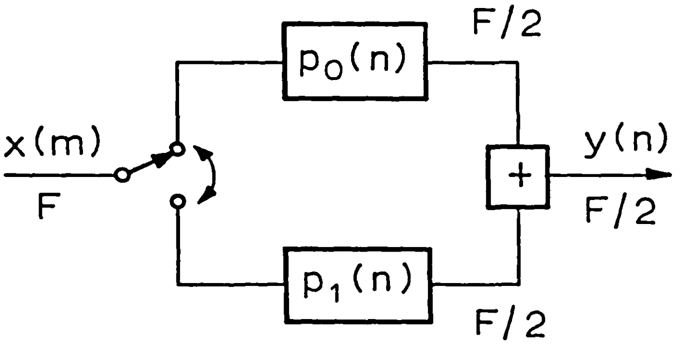


(a) Decimation by two.

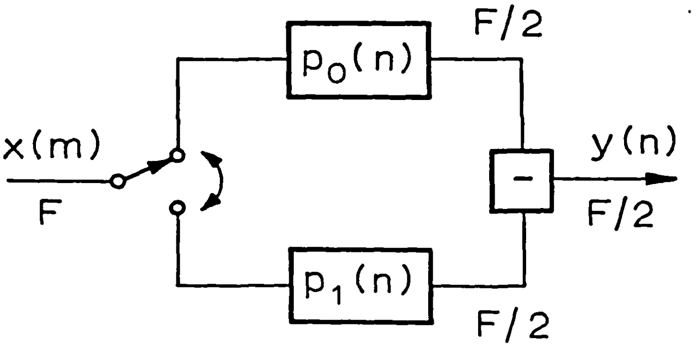


(b) Interpolation by two.

Fig. 3.24 Polyphase structures.

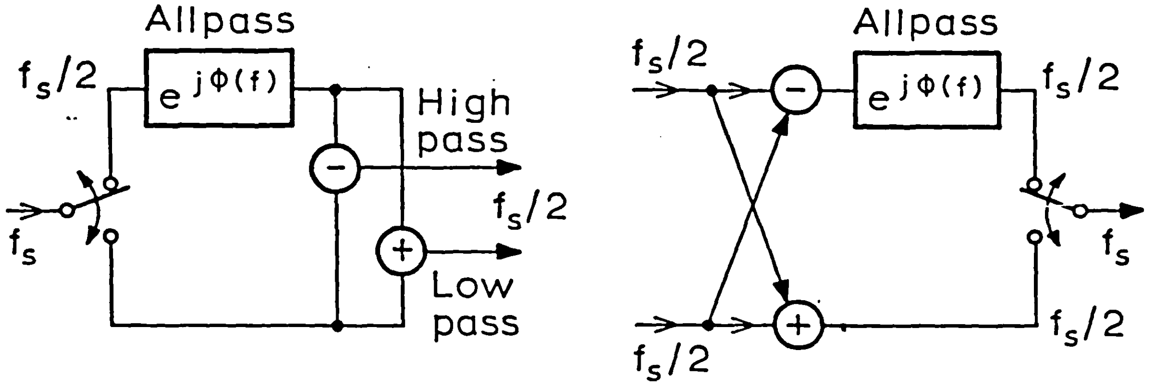


(a) Decimation by two (Lowpass)



(b) Decimation by two (Highpass)

Fig. 3.25. Polyphase filter units.



(a) Decimator (b) Interpolator

Fig.3.26. 1-D allpass units.

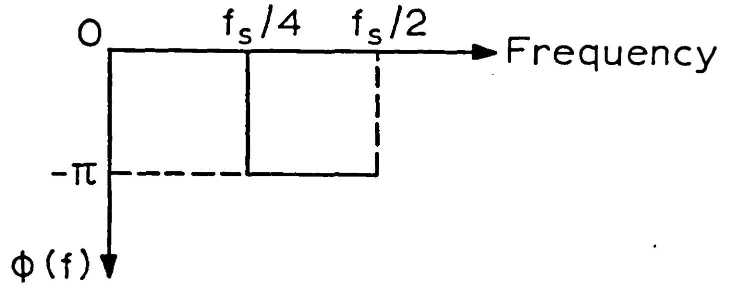


Fig.3.27. Phase-shifter characteristic.

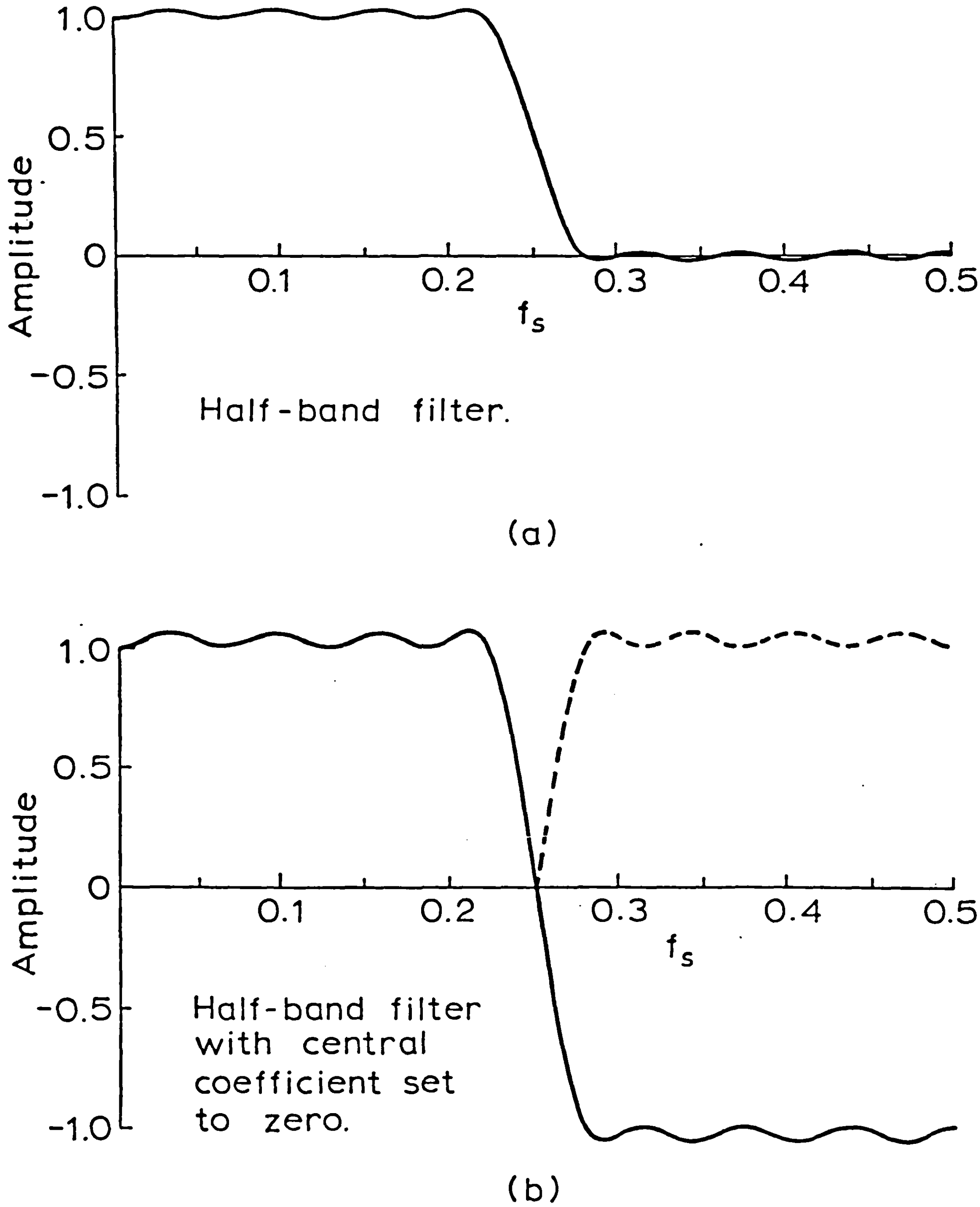


Fig.3.28 1-D allpass units.

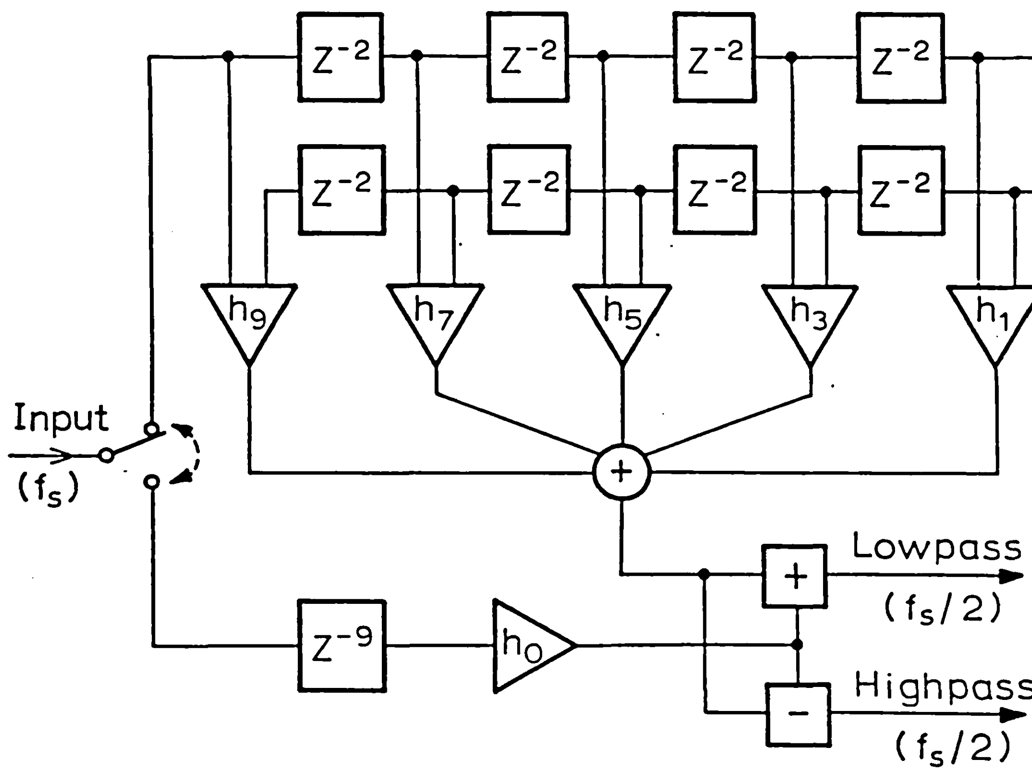


Fig. 3.29 Fir realisation.

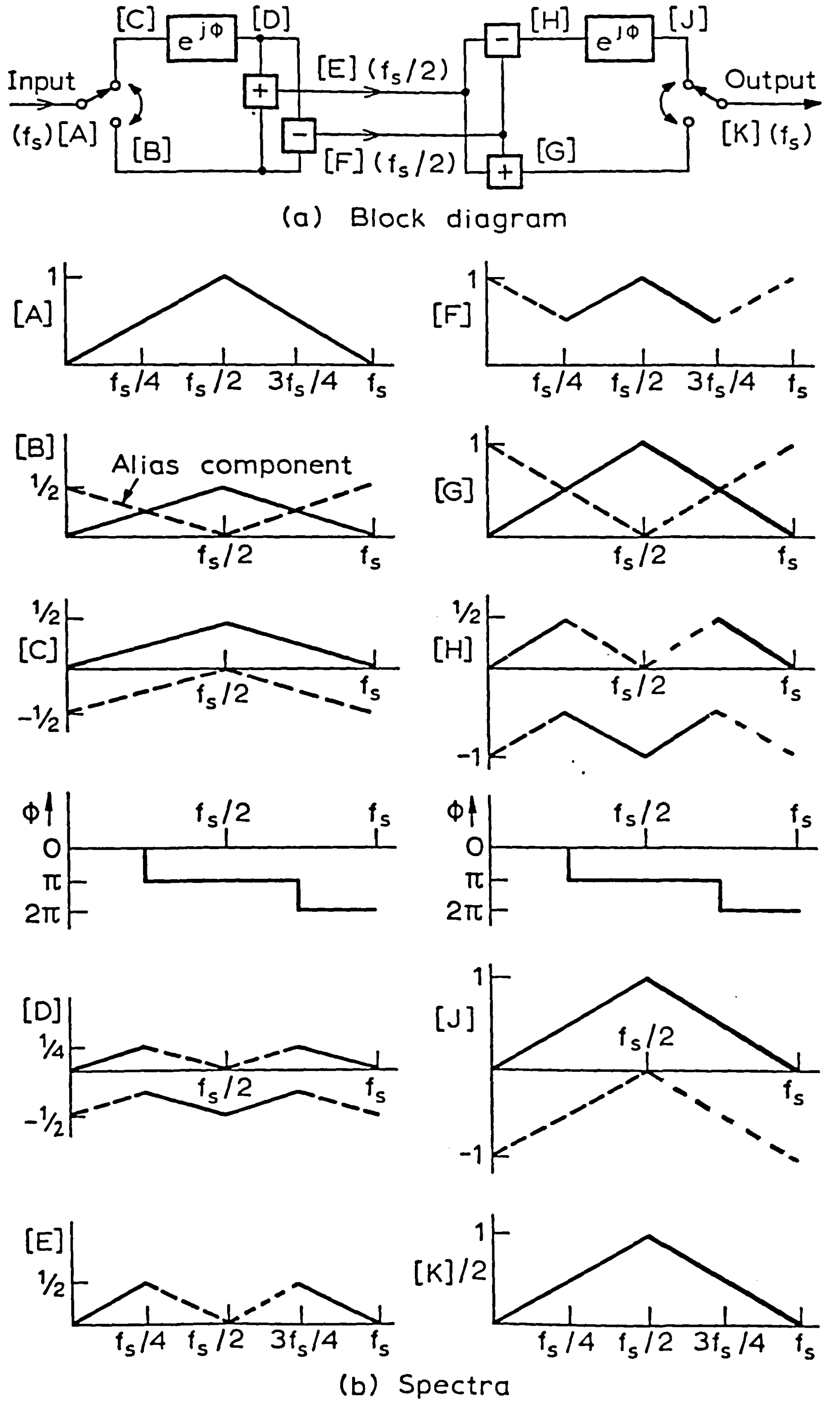


Fig. 3.30 Decimation and interpolation.

CHAPTER 4

COMPUTER SIMULATIONS

4.1 Introduction

The previous chapters have outlined existing techniques for designing multi-dimensional filters. In chapter 3, a new filter design technique was introduced which proved to have numerous advantages over existing techniques. However a common problem with all these techniques has often proved to be the hardware requirements when implementing real-time digital filters. Furthermore, the hardware requirements increase considerably if temporal filtering is required. For example, one frame of a 625-line black-and-white television signal, sampled at the European Broadcasting Union (EBU) standard of 13.5MHz, contains 540,000 pixels. A simple second-order temporal filter would therefore require over a megabyte of digital storage. A colour signal (Y,U,V components sampled at 13.5MHz, 6.75MHz and 6.75MHz respectively) would require twice this amount of storage. Although the new filter structure does not necessarily reduce the amount of hardware required for a specific application, it does allow the hardware to be modular in design and construction, and therefore encourages the use of VLSI.

To investigate fully the new filter design technique, both technically and subjectively, and to enable a full comparison to be carried out with other design techniques, a large and very flexible real-time digital memory would be required. Various hardware implementations were considered for the memory, however they proved to be very bulky and costly. Typically, flexible implementations of nineteenth order horizontal and vertical phase-shifter networks, at video sampling rates of up to 27MHz, would require 2 and 3 double Eurocards respectively if VLSI techniques were not used. Therefore, implementing the filter structure shown in fig.3.10, would require 16 double Eurocards. This requirement could be reduced to 14 boards by exchanging the horizontal and vertical units in fig.3.10 and so produce a more efficient filter structure. Further reductions in the computational hardware (as opposed to the storage hardware) could be achieved by using time-division multiplexing as described in section 3.4.2. However, in the case of temporal filters, (where the storage hardware is relatively large), the saving would be minimal.

Therefore to minimise the cost and complexity of the initial experimental hardware, a decision was taken to utilise a non-real-time computer video simulator facility which was available at the time. This approach also minimised the delay in obtaining initial results on the new filter design technique, and allowed any proposed filter structure to be fully simulated before the real-time hardware or VLSI design was finalised.

A general description of the simulator hardware is given in the next section. Existing software used to design the one-dimensional filter units is also discussed, and some examples of the one-dimensional filter designs used in forming the all-pass polyphase networks are given. The filter simulation software developed for one and two-dimensional filter structures is then be described briefly. Further descriptions of the simulator software are given in Appendix 1. Finally the results obtained from simulating one- and two-dimensional filter structures are discussed. The facilities for processing moving pictures were very limited and therefore the simulations were restricted to two-dimensional filter structures.

(Over 200 separate simulations were carried out on the Digital Video Processing Simulator using different one-dimensional filter designs and two-dimensional filter structures. Only the most important results are discussed in this chapter, however all the results were stored on high-quality video tape and are presently being kept by Philips Research Laboratories).

4.2 Simulation Hardware

Full descriptions of the simulator hardware and software are given in refs. <63> and <64>. A simplified block diagram of the Digital Video Processing Simulator (DVPS) is given in fig.4.1. A Philips P857 minicomputer is interfaced directly to a 2.5 megabyte real-time video memory. The minicomputer is used to control real-time data capture from any video source (eg. off-air, video tape recorder (VTR), or camera) and to control the real-time display of the memory contents on a video monitor. The results may also be recorded on a VTR.

The particular series of video frames required for signal processing is first captured using the real-time memory. The required processing algorithm is then programmed on the minicomputer. The computer can then retrieve the required video samples from the video memory for processing, and write back the processed samples for continuous display on the monitor. In most cases, the computer is not capable of processing the samples at the full video sampling rate and therefore real-time simulation is not possible.

The organisation of the real-time video store is shown in fig.4.2a. The basic memory is arranged as 4096 X 640 pixels. The 12-bit horizontal and 10-bit vertical address generators required to implement the system are combined in two fully programmable 22-bit address generators. The default organisation of the memory is 4 X 1024 X 320 X 2, ie. 4 pictures each with two fields of 320 lines of 1024 horizontal samples. However, the memory may be further sub-divided, both horizontally and vertically, by factors of integer powers of two. For example, if the memory is further sub-divided by 8 in the horizontal direction and 4 in the vertical direction, then 128 pictures, each with 2 fields may be stored, processed and displayed. Each field will then contain 128 samples horizontally and 80 lines vertically. The memory sub-division and the 22-bit address generators are completely under software control and may be changed within the processing algorithm programs.

The particular memory organisation chosen for all the simulations is shown in fig.4.2b. The memory can effectively store 8 pictures (16 fields), each with 512 samples horizontally and 640 lines vertically.

4.3 Design of the One-Dimensional Half-band Filters

We have seen in section 3.6 that an efficient implementation of the new filter design technique requires the design of one-dimensional half-band filters for the one-dimensional allpass phase-shifter networks. Many computer programs exist for the design of such filters, however one particular method is described which produces half-band filter designs where all the filter coefficients are sums of integer powers of two. The complexity of the coefficient multipilers is therefore reduced. A wide range of filter designs was produced to allow investigations into the subjective effects of various filter parameters, such as :

i) Passband and stopband ripple

ii) Transition bandwidth

iii) Transient response

A summary of the results obtained from using various filter parameters is given in this section

4.3.1 Filter design software

Section 3.4.2 mentioned that the efficiency of the one-dimensional filter units could be increased by implementing half-band FIR filters with coefficients of sums of integer powers of two. This restriction on the coefficients minimises the 'shift and add' operations required in the hardware implementation of the coefficient multipliers. Lim and Constantinides <54> have described a linear programming technique which allows the coefficients of FIR filters to be restricted to sums of powers of two. The computer program, ('MILP2'), developed by Lim, is described in ref. <65>. The program allows the user to select one of the following discrete coefficient spaces :

i) single integer power of two

ii) sum of two integer powers of two

iii) sum of three integer powers of two

iv) integer

Fig.4.3 shows the general specification of a lowpass filter. Passband gain is given by 'B' and passband and stopband ripple are given by δ_1, δ_2, and δ. However, section 3.6.1 mentioned that a half-band filter possesses the property :

$$\delta_1 = \delta_2 = \delta \qquad \text{.... (4.1)}$$

The passband (w_p) and stopband (w_s) are defined by :

$$0 \leqslant w_p \leqslant w_1 \qquad w_2 \leqslant w_s \leqslant \pi \qquad \text{.... (4.2)}$$

w_1 and w_2 are determined by the transition bandwidth of the filter, however for a half-band filter (from Equn.3.18 - section 3.6.1) :

$$w_1 = \pi - w_2 \qquad \text{.... (4.3)}$$

MILP2 allows the filter designer to specify the following parameters :

1) 'XMAXQ' = desired value of δ/B. If the value of δ/B of a discrete solution found by MILP2 is such that δ/B < XMAXQ, then the algorithm terminates. To obtain the global optimum solution, XMAXQ is set to 0.

2) 'NTYPE' = filter type (1=lowpass, 2=highpass, 3=bandpass, or 4=bandstop).

3) 'N' (integer) = filter length.

4) 'NPTS'(integer) = total number of frequency grids between $w = 0$ and $w = \pi$ inclusive.

5) 'L1' and 'L2' (integers) = band edges of the passband and stopband. In fig.4.3, w_1 and w_2 correspond to the L1th and L2th points along the w axis respectively.

6) 'JQ2' defines the discrete coefficient space :

JQ2=1 for a single power of two

JQ2=2 for a sum of two powers of two

JQ2=3 for a sum of three powers of two

JQ2=4 for an integer number.

7) 'XJQ2' gives the maximum permissible power of two (for JQ2=1,2,3), or the largest integer grid of the coefficients (for JQ2=4).

8) 'PD' (real) defines the ratio of passband to stopband ripple. For half-band filters PD is set to 1.

9) 'PW' is generally set equal to the required value of δ/B.

The input file for these parameters is arranged as follows :

XMAXQ,
NTYPE,
N, NPTS, L1, L2,
JQ2, XJQ2, PD, PW.

If the specific requirements of a half-band lowpass filter are considered, the parameters for a global optimum solution are :

0,
1,
N, NPTS, L1, (NPTS - L1 + 1),
JQ2, XJQ2, 1, 0.

Equn.3.14 (section 3.6.1) shows that every non-zero even index coefficient of a half-band filter is zero. However, as MILP2 was written to design standard FIR filters, the program will calculate these coefficients, known to be zero. This particular inefficiency may be avoided by defining the order of the filter 'N' to be the number of non-zero coefficients 'Na ' given by :

$$Na = (N+1)/2 + 1 \qquad \text{.... (4.4)}$$

where $N=4M-1$

$$\text{ie. } Na = 2M + 1 \qquad (M - \text{integer}) \qquad \text{.... (4.5)}$$

The changes in the MILP2 software required to support this modification are described by Mpofu <66>.

For example, if a nineteenth-order (N=19) half-band FIR lowpass filter is required, with a minimum stopband rejection of 40dB, a transition bandwidth of 0.23π, and a maximum coefficient size of 255 (8 bits) and the sum of two powers of two, the input parameters would be :

.01,

1

11, 77, 30, 48,

2, 256, 1, 0.1.

The result of giving these parameters to MILP2 is shown in fig.4.4. The coefficients for the filter are :

h(0) = 192

h(1) = 120

h(3) = -36

h(5) = 16

h(7) = -8

h(9) = 3

giving a scaling factor for unity passband gain of 1/384.

MIPL2 gave the stopband rejection as -42.098 dB. If the central coefficient (h(0)) is removed, the frequency response shown in fig.4.5 is obtained. Scaling the response by two will give the allpass phase-shift response required (section 3.6.2).

Mpofu <66> also carried out some interesting studies regarding the effect of constraining the coefficients to discrete coefficient spaces. In general, his results show that for the best combination of resolution and coefficient (ie computational) complexity, the coefficients should be restricted to the sum of two-powers-of-two. Most of the filters designed for the simulations were specified to have coefficients of up to two-powers-of-two and 255 (8-bits).

4.3.2 Subjective assessment

Extensive subjective tests were made with many half-band filter designs in both the one-dimensional and two-dimensional filters structures. The aim of the tests was to find the minimum filter order required for an 'acceptable' picture quality and to gather information on the relative subjective effects of filter parameters such as passband and stopband ripple, transition bandwidth and transient response.

In most cases the minimum filter requirement for broadcast quality pictures was a 15th order filter with a passband ripple of 0.25dB and a transition bandwidth of 1.5MHz, given the criteria that filter artefacts such as passband ripple and alias component rejection could not be detected at a viewing distance of 5ft on a high-quality (Barco) 26" colour monitor.

4.4 Simulation software

All the simulator programs were written in FORTRAN IV and are described briefly in Appendix 1. The analogue-to-digital convertor at the input to the simulator was set up in software to sample the incoming black and white video signal at the EBU standard of 13.5MHz for luminance. The real-time video memory was set up to store 8 pictures (16 fields) of 512 horizontal samples and 640 lines. The standard 625-line television has 575 lines of active video and about 700 active pixels per line if sampled at 13.5MHz. Therefore the effective picture size shown in all the results is 512 X 575 pixels.

If the lowpass component in any of the filter structures is switched out, the dc pedestal in the video signal will also be eliminated from the output. Therefore, to allow correct decoding of the line and field syncronisation signal at the monitor, the dc pedestal must be restored by the software. Evidence of this process may be detected in some of the photographs as a slight overall increase or decrease in the brightness level, due to the incorrect dc level being chosen by the software.

4.5 Simulations of One-dimensional Filter Units

The first simulations were carried out to investigate the operation of the basic one-dimensional filter units shown in figs.3.26 and 3.30. Program UNIT simulates the cascade of a horizontal decimator unit and a horizontal interpolator unit.

Fig.4.6 shows a horizontal multiburst with horizontal frequencies of 1 to 6MHz in 1MHz steps. The original analogue signal has been digitised by the simulator at 13.5MHz and D/A converted without any processing by the host computer. A noticeable fall off in the frequency response of the simulator and pattern generator may be seen in the 6MHz burst.

Fig.4.7 shows the results of simulations for the lowpass, allpass, and highpass configurations, using the phase shift filter of fig.4.4 and a sampling rate of 13.5MHz. Referring to the block diagram of the cascaded decimation and interpolation units in fig.3.30a, the outputs were obtained with the following signal combinations at the centre of the filter structure :

Lowpass : [E] only

Highpass : [F] only

Allpass : [E] and [F]

Fig.4.7 confirms that the cascade structure of a decimator unit and an interpolator unit is performing as expected, with the transition between the lowpass and highpass outputs occuring at $f_s/4$ = 3.375 MHz. There is no noticeable difference between the allpass version and the original signal. However, if the multiburst had been continuous rather than discrete, some distortion of the signal would have been visible around the transition frequency, due to the imperfect allpass response of the phase-shifters. As expected, the black level in the highpass output is different to that in the lowpass output, due the computer having to estimate the dc component lost during highpass filtering.

4.6 Simulations of Two-dimensional Filter Structures

Two software packages were developed to simulate the basic two-dimensional filter structure shown in fig.4.8a. Both packages allow a choice of four sections in frequency space and they differ only in the choice of effective vertical sampling rate at the input to the filter structure. The first software package, 'STAG2', has an effective vertical sampling rate of 312cph (fig.4.8b) and this was achieved in the simulation software by processing each field (312 lines) of video in turn. The second package, 'STAG3', shows the vertical sampling rate as 625cph (fig.4.8c) and this was achieved in the simulation software by processing each frame (625 lines) of video in turn.

Fig.4.9 shows a two-dimensional 'Zone-Plate' <67> after A/D and D/A conversion and no processing by the host computer. This particular zone-plate contains a two-dimensional frequency sweep of 0-6MHz along the horizontal axis and 0-156 cph along the vertical axis. It therefore shows, directly, the two-dimensional frequency response of the system.

Fig.4.10 shows the output of the simulator using the Zone-Plate of fig.4.9. The STAG2 software was used to keep frequency sections A, B, and C of fig.4.8b and reject frequencies in section D. Fig.4.11 shows the theoretical frequency response of the filter structure. Again the filter output agrees well with the theoretical response. Some distortion is visible at both the horizontal and vertical transition frequencies (3.375MHz and 78cph respectively) due to the approximate allpass response of the phase-shifters. Some aliasing due to the line scanning may also be detected in both the original and filtered outputs at around 156cph. The effect of setting the initial conditions for the filter to black-level may be seen at the edges of the picture. The one-dimensional half-band filter used to obtain the phase-shift response in software was designed using MILP2 with the following characteristics:

i) Filter Order = 31

ii) Passband Ripple = 0.5 dB

iii) Transition Bandwidth = 284 KHz

Fig.4.12 shows another video input signal after A/D and D/A conversion and no processing by the host computer. Areas of interest in the picture include the horizontal, vertical and diagonal frequencies in the shirt, and the diagonal frequencies in the jackets.

Fig.4.13 shows the effect of using the STAG2 software to switch in sections B,C and D of fig.4.8b and switching out section A. Only high spatial frequencies are now visible in the picture. The characteristics of the half-band filter were as follows:

i) Filter Order = 31
ii) Passband Ripple = 0.5 dB
iii) Transition Bandwidth = 284 KHz

Fig.4.14 shows the effect of processing the picture using the STAG3 software with sections A,B and C of fig.4.8c switched in, and section D switched out. Some loss of resolution in the high diagonal frequencies may be seen. Again the effect of incorrect initial conditions may be seen at the edges of the picture. The half-band filters had the following characteristics:

i) Filter Order = 19
ii) Passband Ripple = 0.1 dB
iii) Transition Bandwidth = 556 KHz

Fig.4.15 shows the effect of field quincunx downsampling on the original picture (fig.4.12). (Fig.4.16a shows the exact form of the field-quincunx downsampling structure). Distortion in diagonal frequencies is clearly visible. Fig.4.16b shows the theoretical spatial filter response required for the pre- and post filters when using this particular downsampling structure.

Figs.4.17 and 4.18 show the effect of pre- and post-filtering the downsampled picture using STAG2 and STAG3 software respectively. Both filter structures kept frequency sections A,B and C and rejected section D, and both used the following design of half-band filter:

i) Filter Order = 19
ii) Passband Ripple = 0.1 dB
iii) Transition Band width = 556 KHz

Although the filter response is a poor approximation to the desired response for field-quincunx downsampling, most of the distortion in diagonal frequencies has been eliminated by the pre- and post-filtering. The picture filtered by STAG3 is generally better than that filtered by STAG2 because the frequency response is closer to the theoretical response required (fig.4.16b).

4.7 Summary

A large amount of hardware would be required to implement a flexible real-time filter structure capable of fully testing and assessing the filter structures both theoretically and subjectively. Various filter structures were therefore simulated in non real-time, using a computer controlled Digital Video Processing Simulator. The system hardware and software associated with the simulator have been briefly described, and a brief summary of the particular filter structure programs is given in Appendix 1.

The half-band filters required for the phase-shifter units were designed using an existing filter design package by LIM <54>. The criterion for high-quality pictures was no noticeable degradation (due to the ripple, transient response etc. associated with the phase-shifters) on a high-quality 26" monitor colour at a viewing distance of 5ft. The phase-shifter characteristics required to meet this criterion were approximately :

N = 15

Passband Ripple = 0.25dB

Transition Bandwidth = 1.5MHz.

Numerous one- and two-dimensional filter structures were simulated, and some of the results have been discussed. The results were encouraging and proved the effectiveness and flexibility of the filter structure. However, the two-dimensional simulations for field-quincunx downsampling indicate that the filter structure may need a resolution of around 16 frequency blocks for acceptable picture quality. This would require an 8 stage filter (4 decimation stages and 4 interpolation stages) and therefore around 120K bytes of storage (using 15th order phase-shifters) for each filter. Although this amount of memory may be obtained on a single chip, it is still significantly (around 5 times) more than that required by a dedicated filter (cf Tonge <28>). The filter structure may therefore prove to be most beneficial in systems which require flexible real-time adaptive filters, such as those utilising movement detectors.

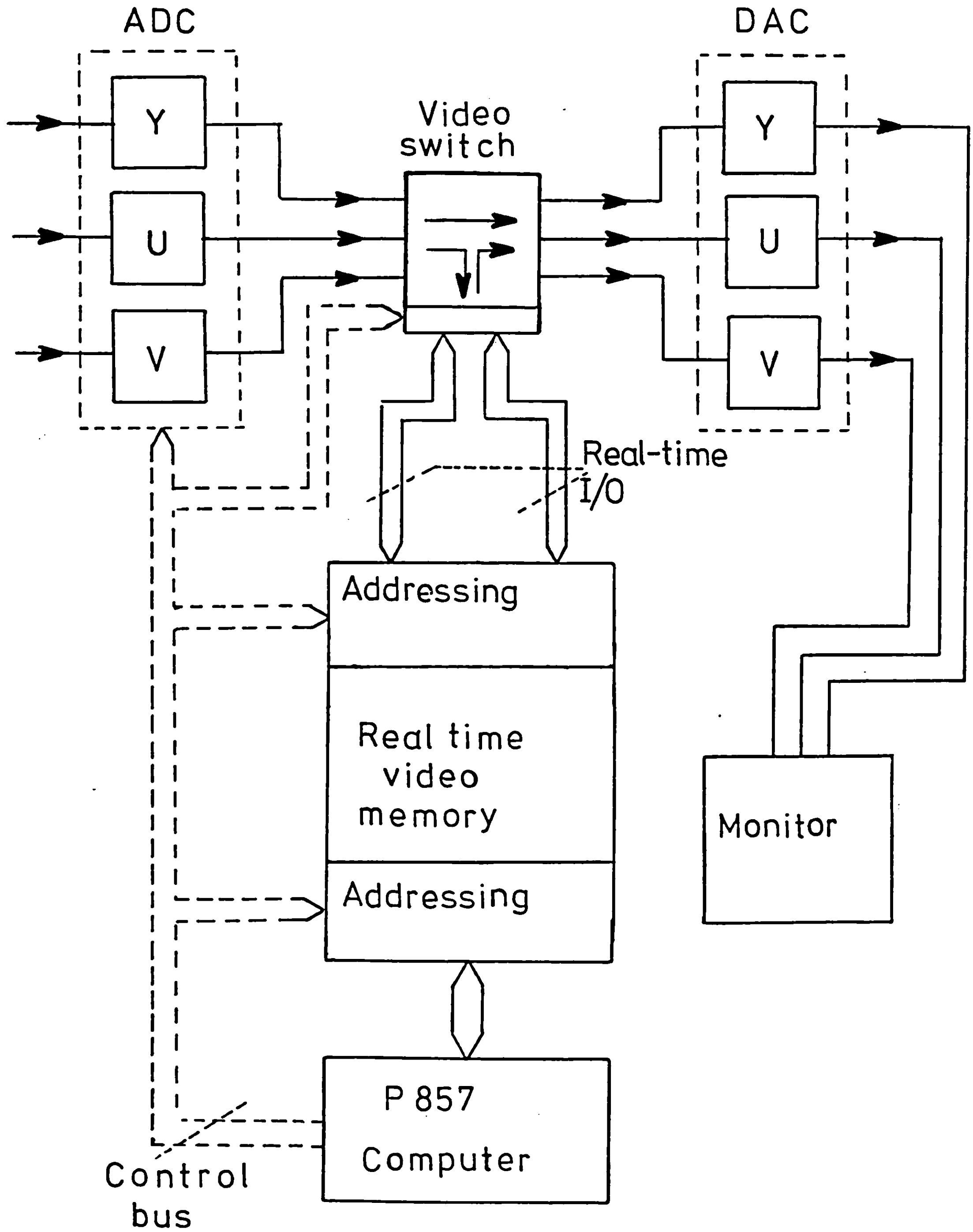


Fig. 4.1 Digital video processing simulator .

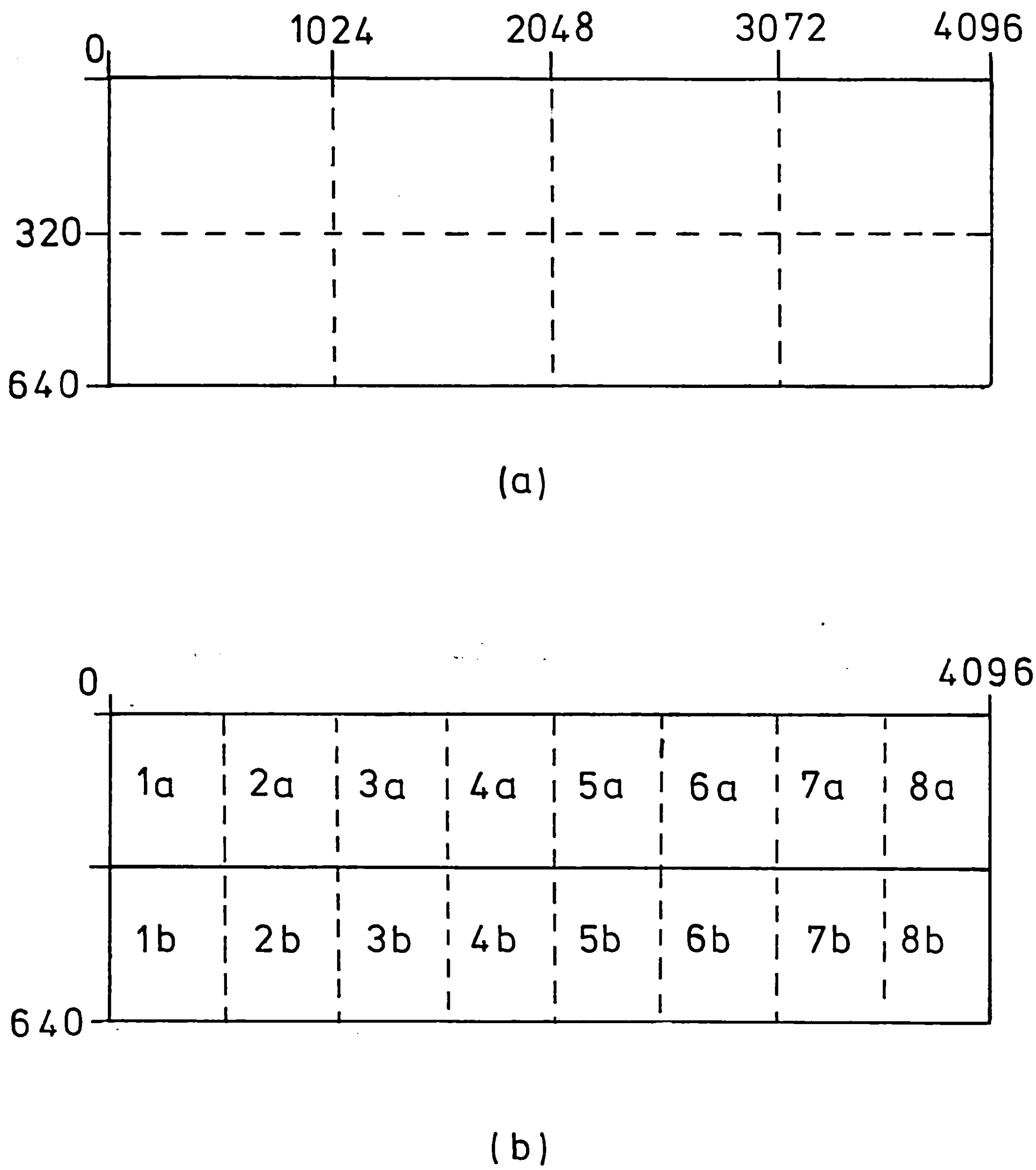


Fig. 4.2 Video memory organisation.

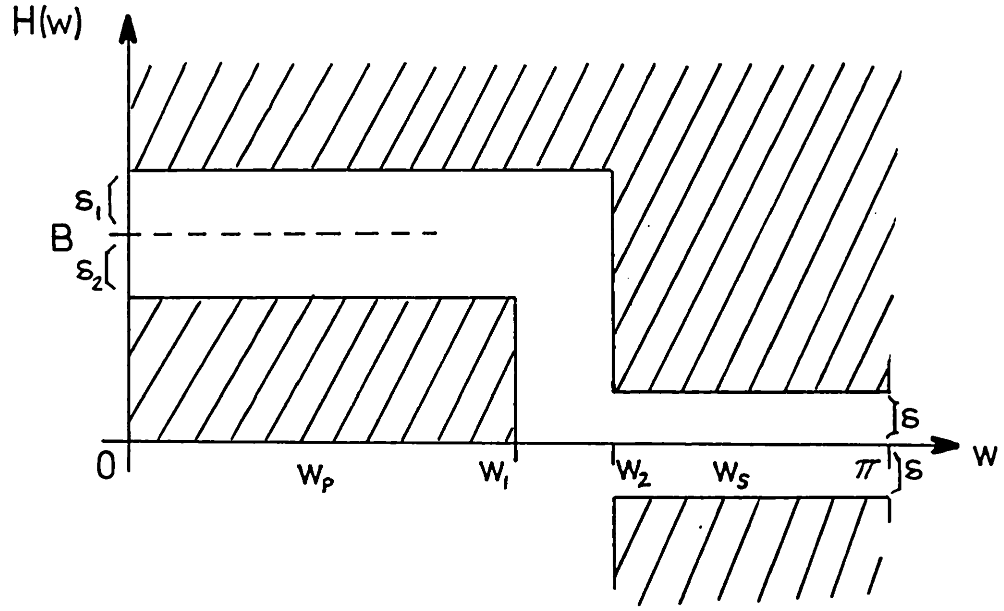


Fig. 4.3 Lowpass filter - general specification.

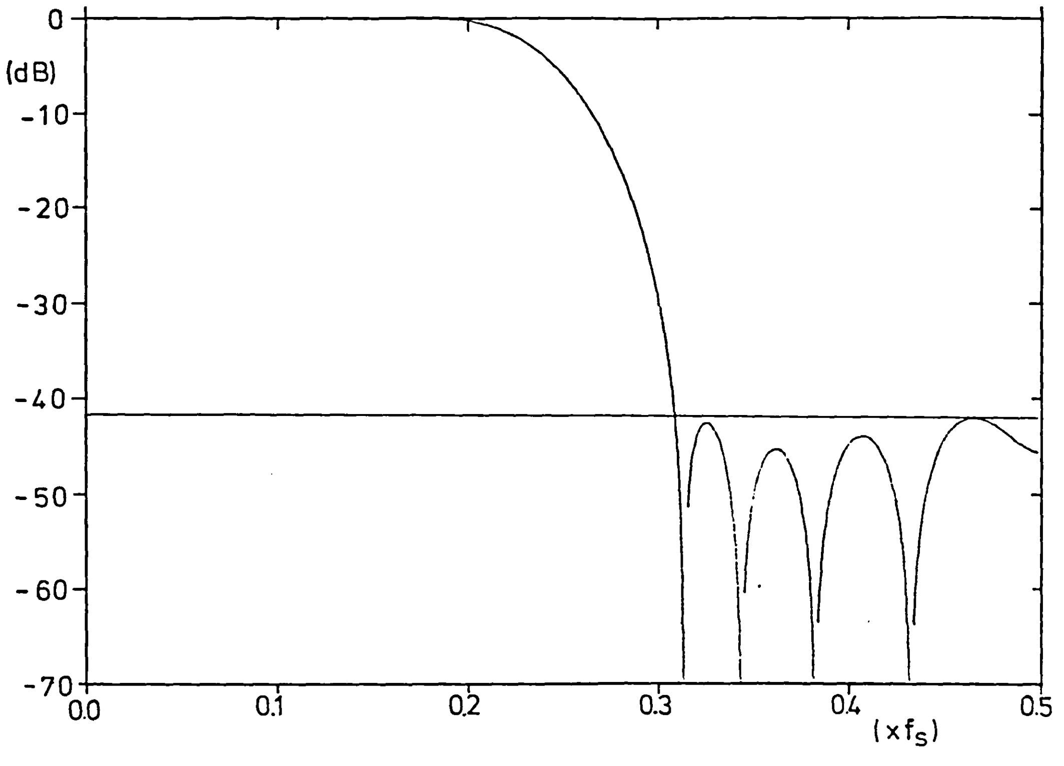


Fig. 4.4 Half band-filter.

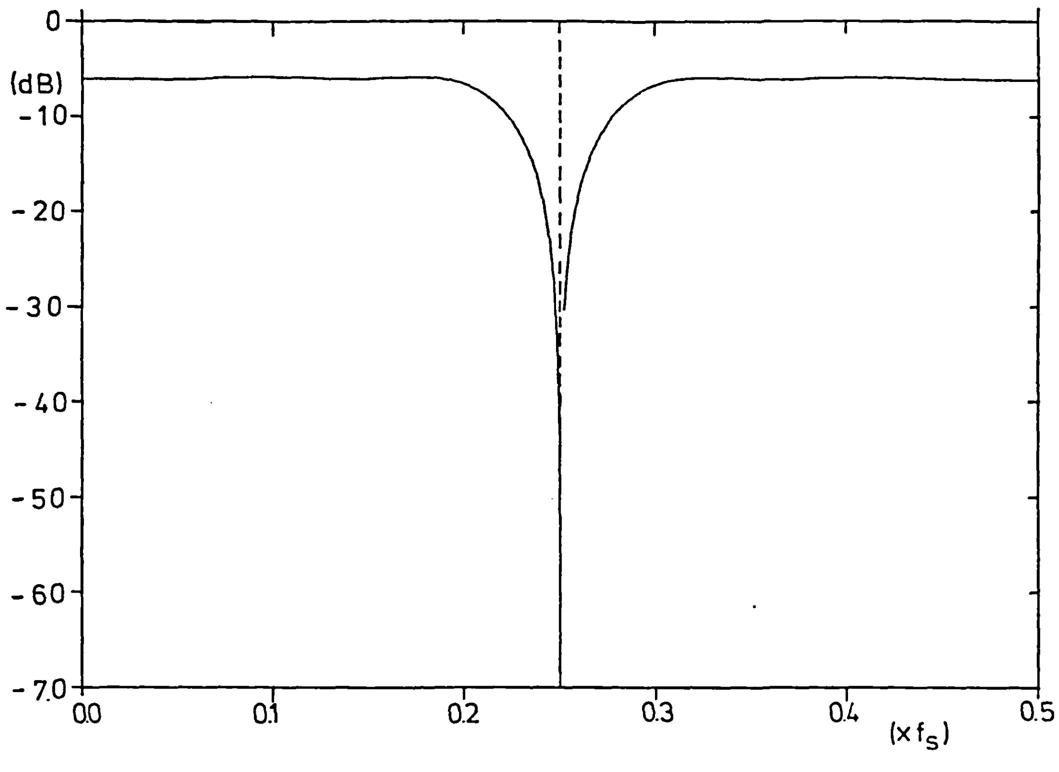


Fig.4.5 Half-band filter with dc coefficient removed.

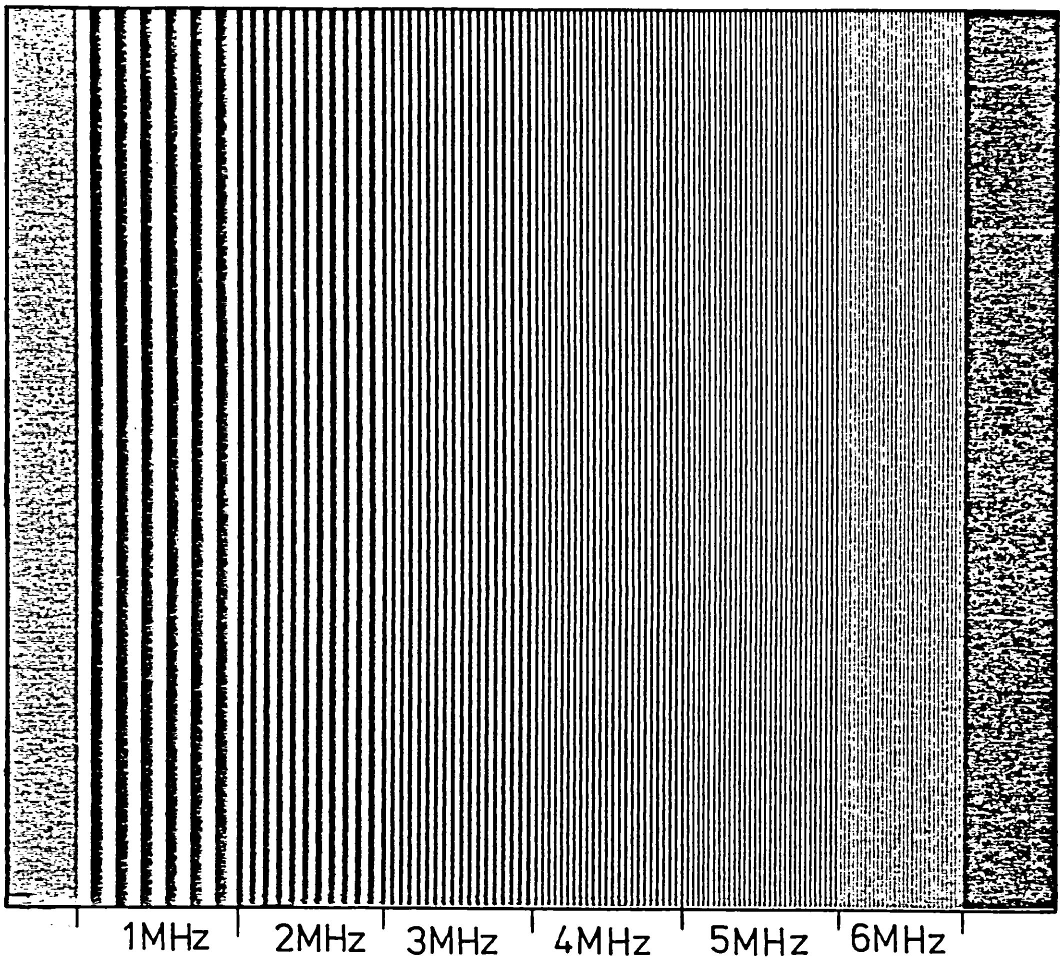


Fig. 4.6 Multiburst - original .

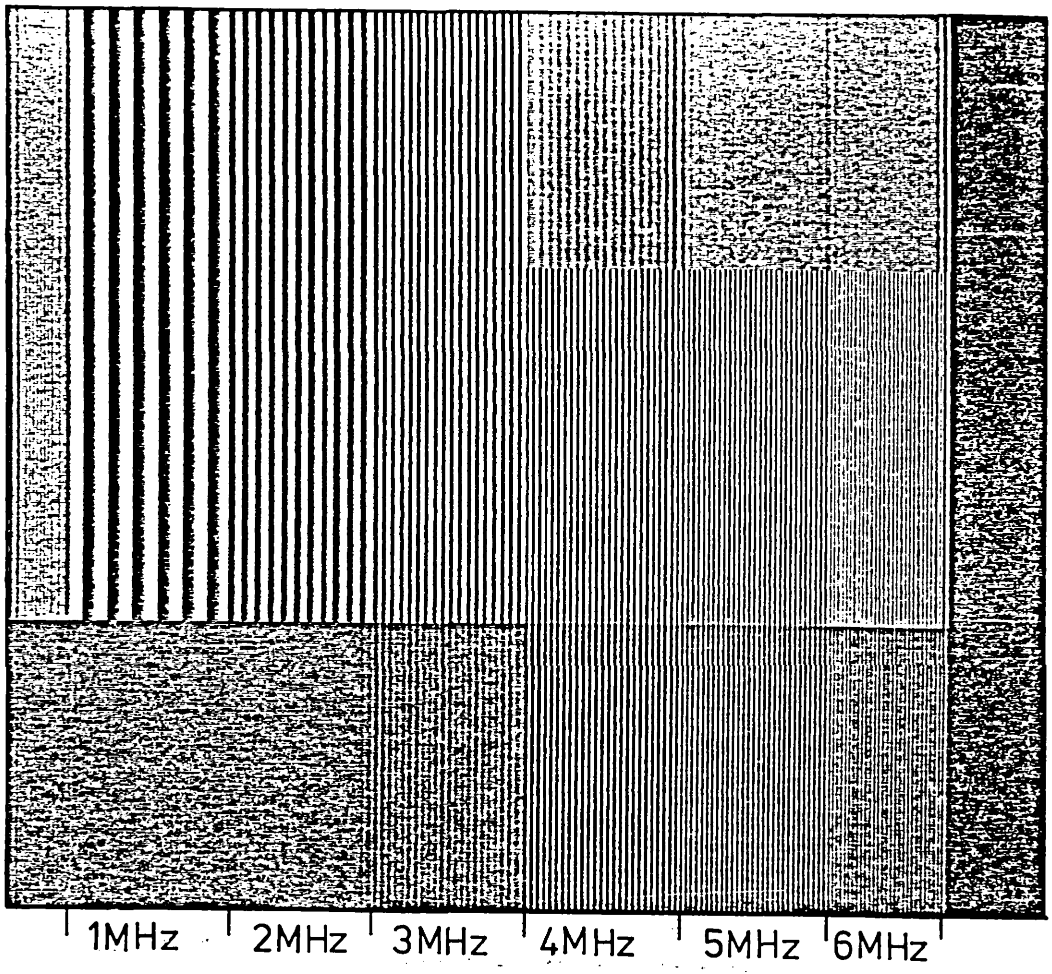
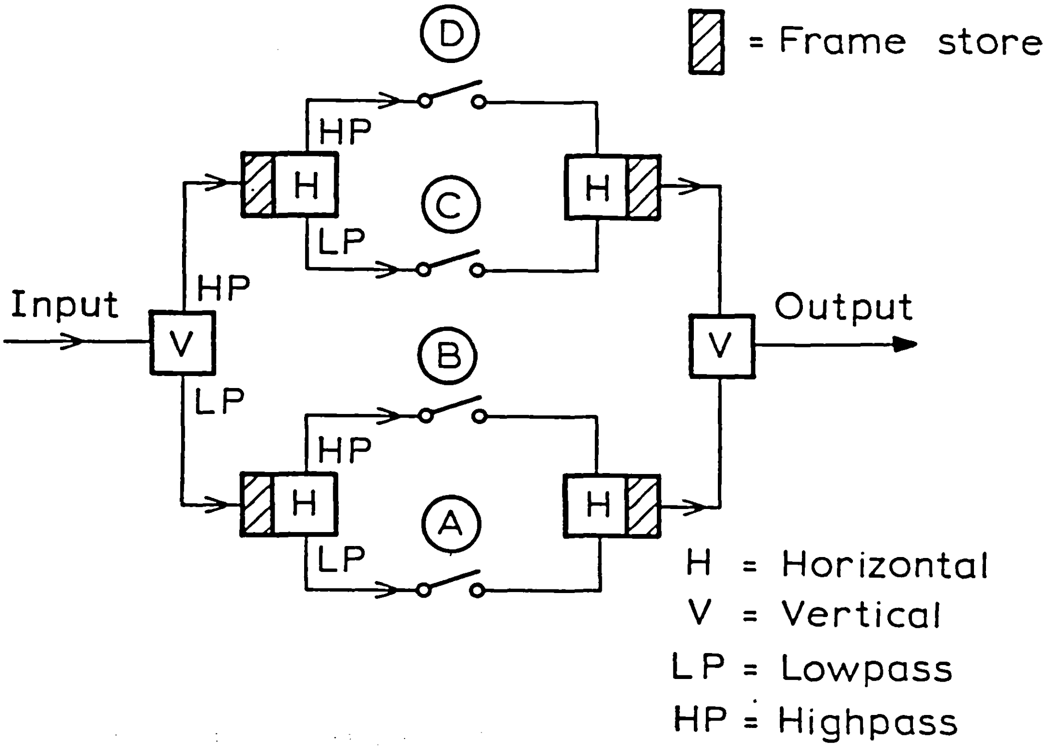
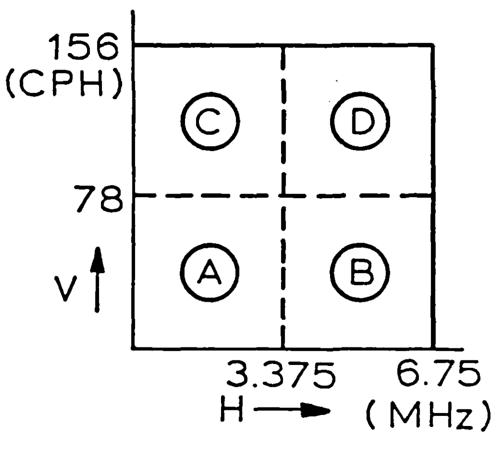


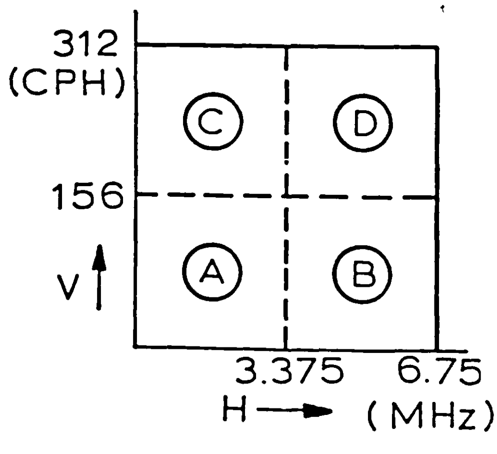
Fig. 4.7 Multiburst - filtered by UNIT



(a)



(b) STAG 2



(c) STAG 3

Fig. 4.8 Simulated 2-D filter structure.

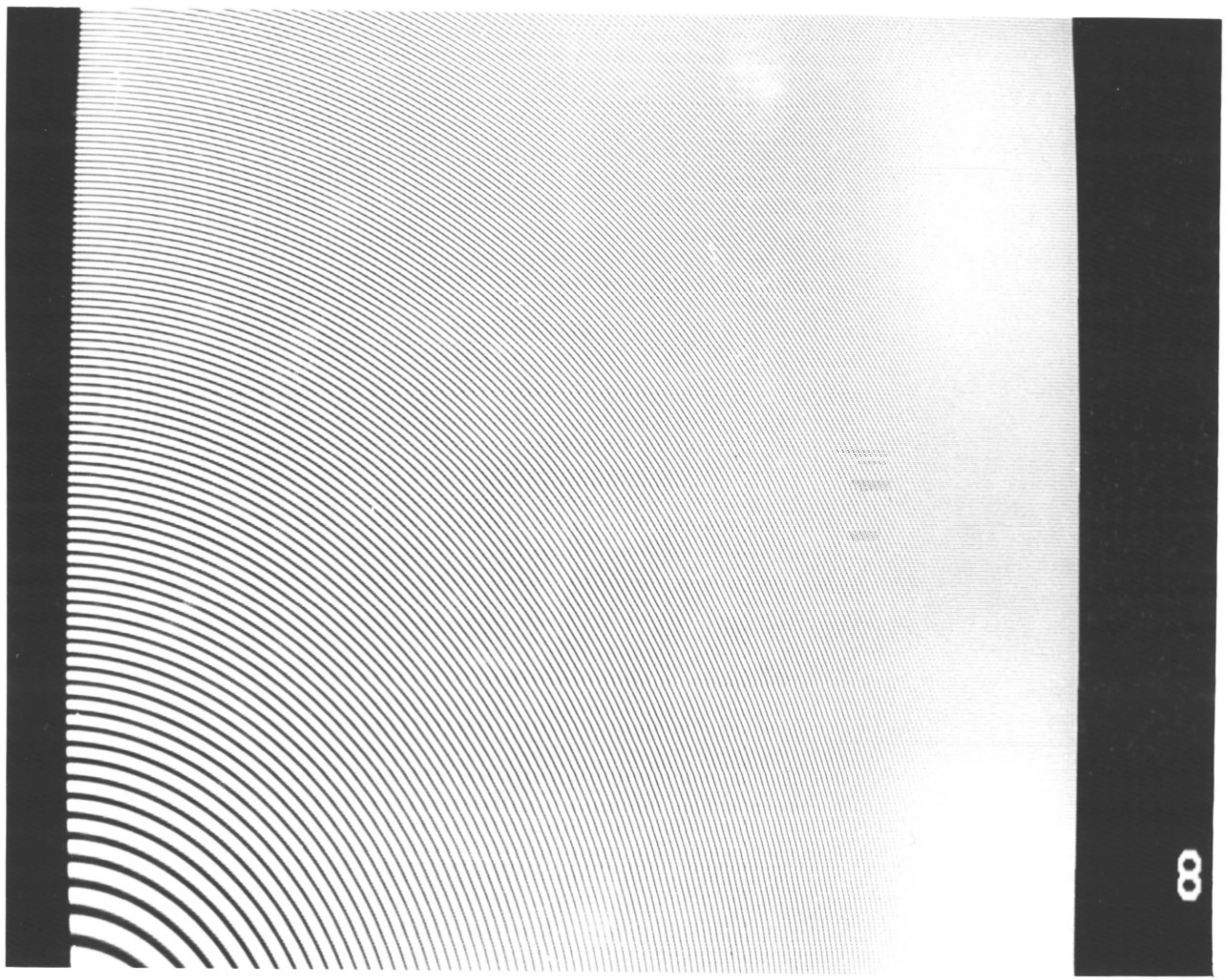


Fig. 4.9 Zone plate - original.

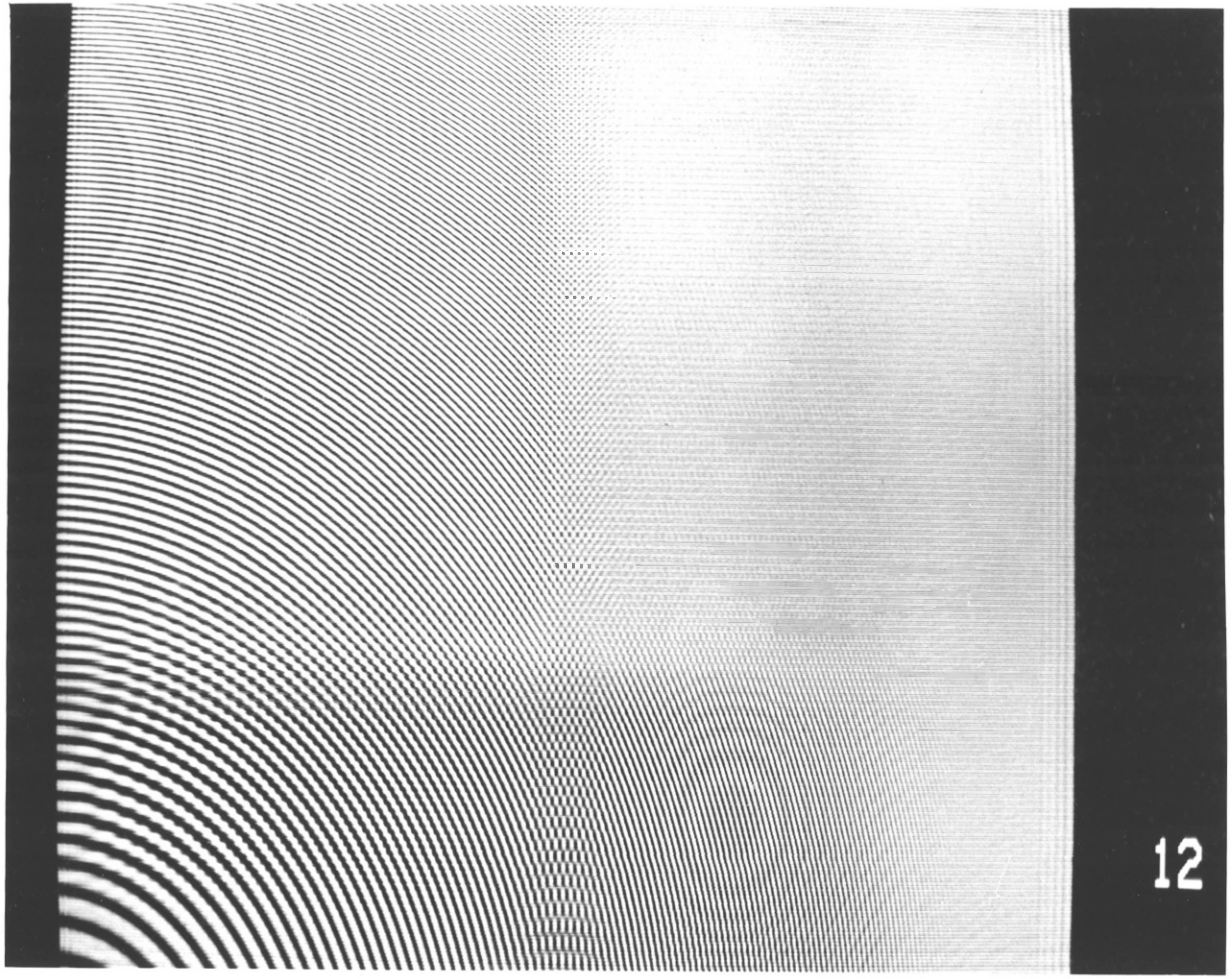


Fig. 4.10 Zone plate - filtered (A,B,C).

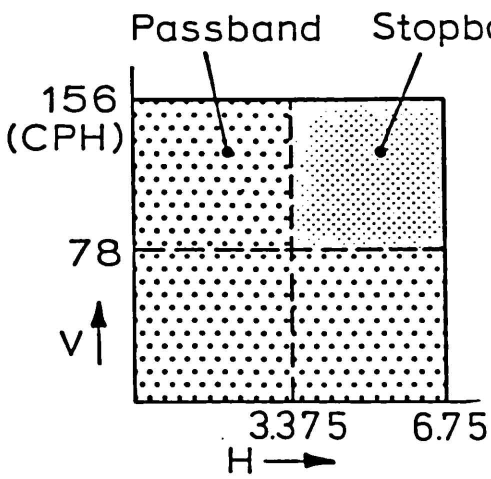


Fig. 4.11 Simulated 2-D filter (STAG 2)

Fig. 4.12 Original.

Fig. 4.13 Filtered (B,C,D).

Fig. 4.14 Filtered (A,B,C) .

Fig. 4.15 Field quincunx down-sampling.

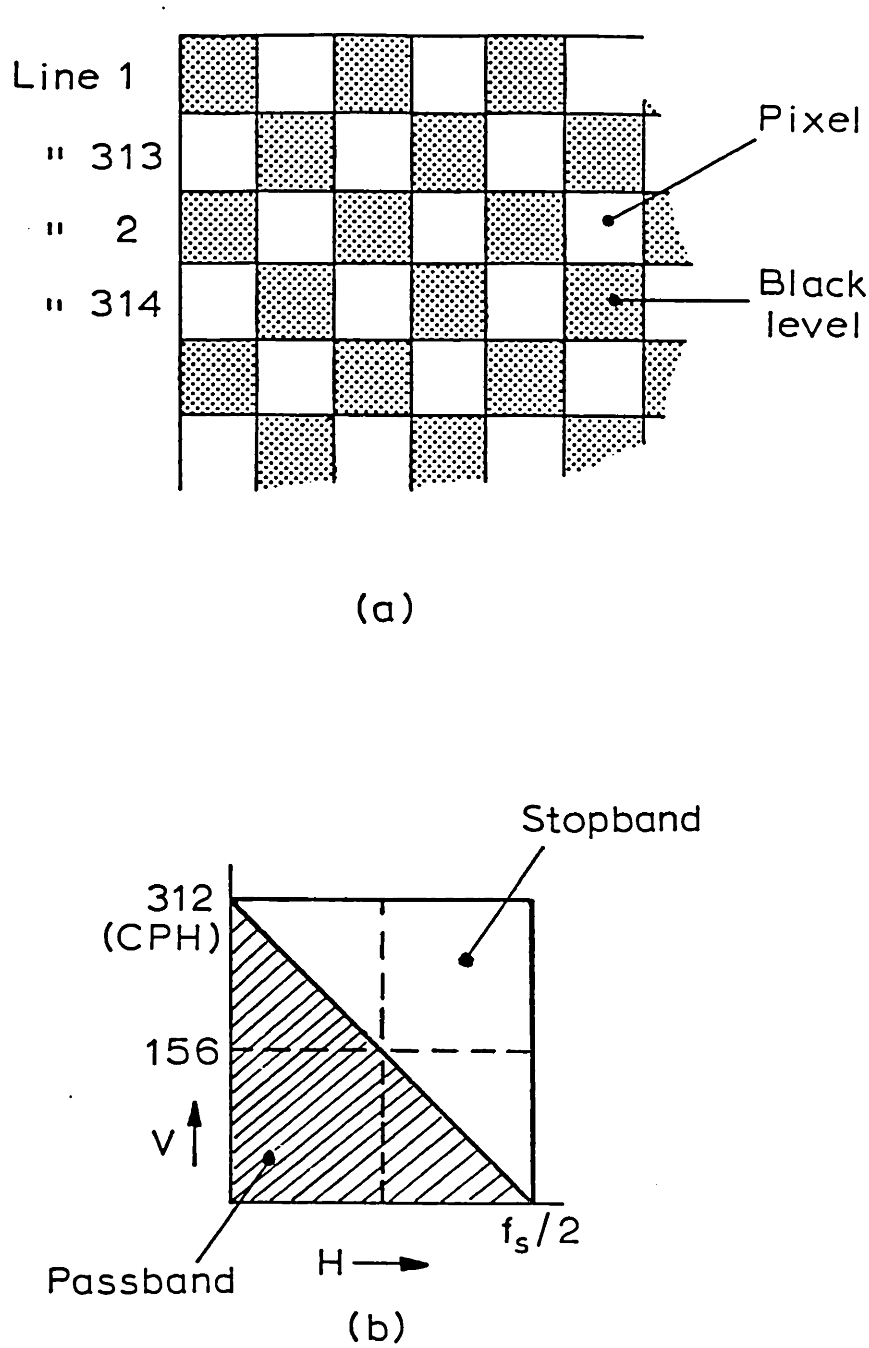


Fig. 4.16 3-D field quincunx down sampling.

Fig. 4.17 Pre- and post-filtering with STAG2.

Fig. 4.18 Pre- and post-filtering with STAG3.

CHAPTER 5

CONCLUSIONS AND FURTHER WORK

5.1 Conclusions

World-wide interest in high-quality television systems has been growing steadily for over a decade, and the recent allocation of satellite broadcasting channels for television has accelerated this interest. The realisation of a high-quality television system with the technical quality of 35mm film and the subjective quality of the cinema is proving to be an attractive proposition to many film and broadcast television companies.

One important effect of this has been the increasing interest shown in the digital processing of television signals. Multi-dimensional filtering and sampling techniques are being investigated with the aim of maximising the subjective quality of the final image within a given transmission bandwidth. However, the design of digital filters for applications in the broadcast industry has concentrated on relatively simple one-dimensional, or separable two-dimensional filters.

It has been shown that a digital television is an example of a three-dimensional sampling system. Various three-dimensional sampling structures have been discussed, together with the associated pre- and post filters. Although knowledge of the three-dimensional bandwidth of the eye can prove useful in the theoretical design of these filters, their evaluation must in the end be subjective.

The idea of redundancy in a video system has been discussed and its elimination may be achieved by tayloring the bandwidth of the system as closely as possible to that of the eye. However, the eye's ability to adapt to viewing conditions and subject matter makes the development of an accurate analytical model of the eye very complex.

To allow a full investigation into sampling structures for future television systems, a new flexible filter design technique has been developed. The technique has two very significant advantages - flexibility, and ease of design. The technique greatly simplifies the design and realisation of flexible real-time multi-dimensional pre- and post-filters.

Flexibility and simplicity are achieved by implementing the filter as a tree-structure made up of one-dimensional filters. With this implementation, it has been shown that non-separable multi-dimensional filter responses may be obtained from a structure of one-dimensional filter units. The major part of the filter design process consists only of using established one-dimensional filter design algorithms.

In common with most multi-filter operations, the technique does not always make the optimum use of hardware when implementing a dedicated filter characteristic. It is relatively easy to minimise the computational hardware employed in the filter structure, and various minimisation techniques have been discussed. However, the amount of storage hardware could prove to be prohibitive if a high filter resolution is required in the temporal direction. There is a direct trade-off between the amount of hardware required and the flexibility of the filter structure, and in dedicated filtering applications the hardware can be reduced. In practice, the modularity of the filter structure also simplifies the hardware implementation and encourages the use of VLSI.

Results obtained from computer simulations of the new filter structure show that the technique is easily implemented and very effective. It has proved invaluable in the subjective analysis of various sampling and filter techniques and the results of some of these investigations have been included. However, as expected, a relatively large amount of hardware is required to implement a real-time filter with comparable performance to that of a dedicated filter. The filter structure will therefore prove most beneficial in systems requiring flexible filters. For example, the filter structure could prove to be very efficient in real-time motion adaptive systems where the stepped three-dimensional response of the filters removes the need for movement detectors.

The flexibility and simplicity of the design method allows many filter characteristics to be obtained from the same hardware with the mimimum of software programming. Real-time control of the filter response is therefore very easy to achieve with a micro-processor or mini-computer, even though the filter hardware itself is operating at sampling rates well above that of the computer.

5.2 Further Work

The study of video sampling structures and filter design techniques will continue to be of great importance to the video industry. A considerable amount of work involving the subjective analysis of the human eye has still to be carried out before the full potential of any image processing system can be realised.

Although the half-band filter tree-structure implementation given in the text is flexible and modular, it is not the only implementation of a multi-dimensional filter structure which uses one-dimensional filters. Two interesting alternatives exist.

One method utilises bilinear transforms and half-band lowpass filtering. A specific example, for use in two-dimensional field-quincunx downsampling, is shown in fig.5.1. The example uses a two-dimensional bilinear transform of the form :

$$H(n,m) = (-1)^{n+m} \qquad n,m \text{ integers} \qquad \ldots.(5.1)$$

to effectively rotate the frequency spectrum of the input signal through 180^{o} in the two-dimensional frequency domain. The process could be thought of as a cascade of two one-dimensional lowpass-to-highpass transformations of the form :

$$H(n) = (-1)^{n}$$
$$H(m) = (-1)^{m} \qquad n,m \text{ integers} \qquad \ldots.(5.2)$$

One-dimensional half-band filters are then cascaded to eliminate the unwanted parts of the frequency spectrum, (in this case the section numbered '4'). Downsampling at the output of each filter would allow further stages to be added to achieve the desired response and resolution.

In general, a multi-dimensional transform of this type, followed by a cascade of one-dimensional filters and down-sampling by a factor of two, will allow the same filters to be realised as with the original tree-structure. Allowing the transformations and filter cascades to be programmable will increase flexibility (fig.5. 2).

This particular method would prove advantageous when realising certain specific filter characteristics, however the reduction in hardware requirement, (over say the structure in fig. 3.11), has been achieved at the expense of a reduction in the overall flexibility of the structure. This alternative structure requires further theoretical and practical investigation.

The filter structures developed so far allow the designer to chose specific areas of passband and stopband in the filter response. No areas of finite attenuation between the passband and stopband levels can be specified. This could prove to be a limitation in certain applications and ways of modifiying the filter structure could be investigated further.

Further theoretical study of the phase-shifter networks could be made to consider the effects of transition bandwidth and passband ripple. Only one specific implementation of the phase-shifters was considered and other implementations could be investigated including the possibility of using IIR filter designs.

The modular hardware implementations which the tree-structure technique encourages, opens up the possibility of VLSI realisation. The tree-structure implementation, in particular, could be implemented using exactly the same VLSI chips for the computations and other VLSI memory chips for the digital filter storage.

The author hopes that further investigation and development will take place on the filtering techniques described in this thesis.

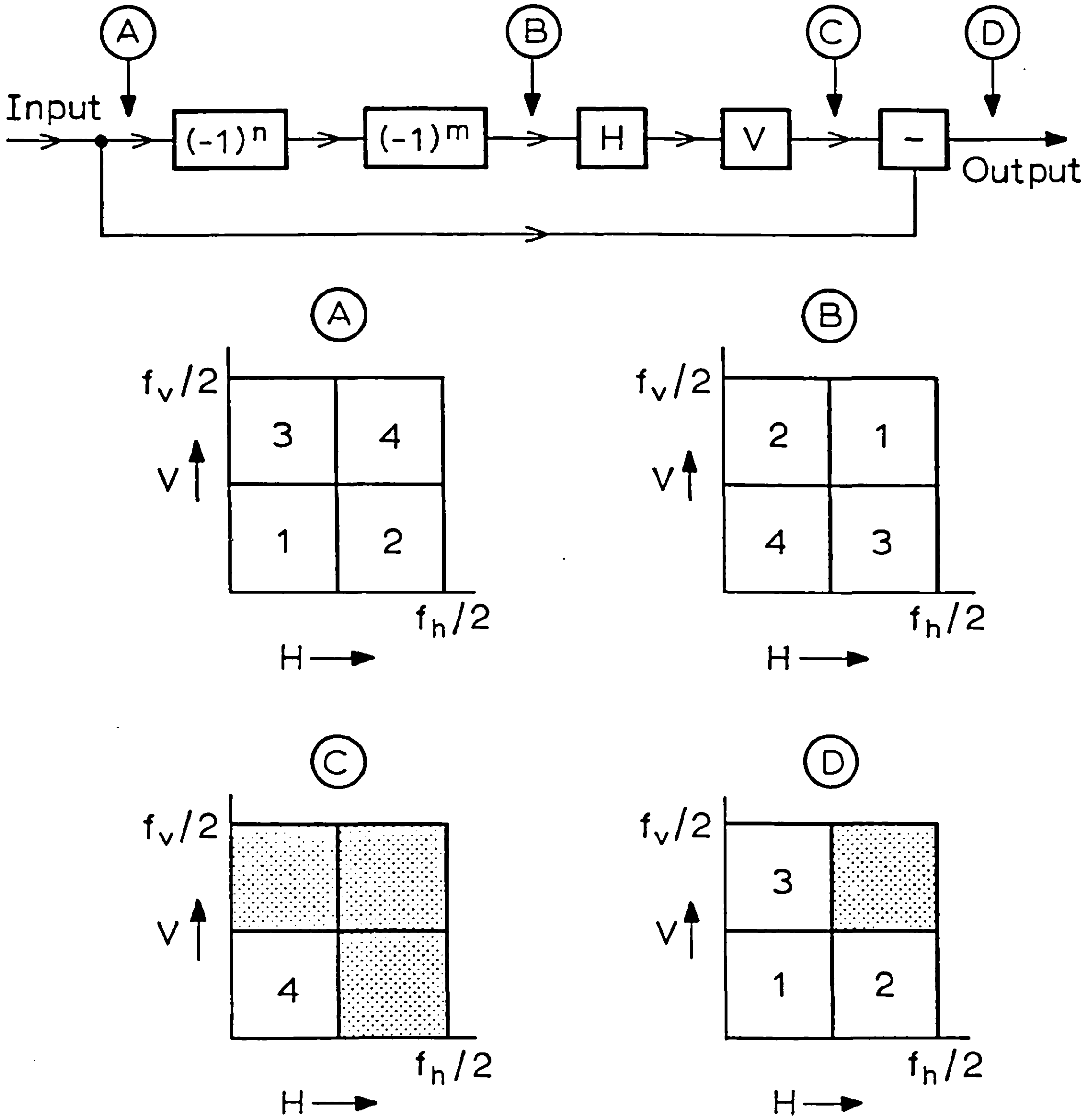


Fig. 5.1 Alternative filter structure.

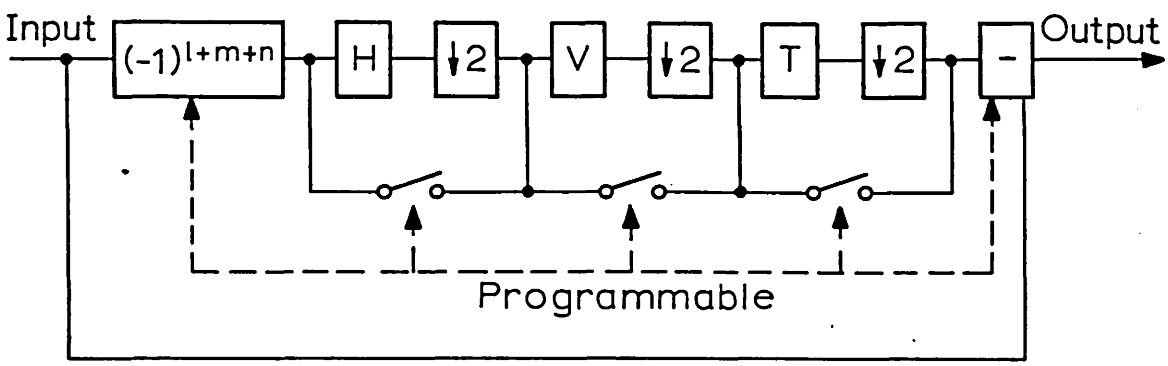


Fig. 5.2. 3-D filter structure.

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APPENDIX 1

SIMULATION SOFTWARE

A1. Introduction

This section gives a brief description of the simulation programs which were written on the Philips P857 minicomputer to investigate the tree-structure filters. All the programs are written in FORTRAN. The reader may find it helpful to read the system software documentation for the Philips Digital Video Processing Simulator <64>. Full descriptions and listings of all the programs are held by Philips Research Laboratories, Redhill, Surrey.

All the simulation programs assume a monochrome video signal with line-locked sampling at between 8Mhz and 16MHz. (In practice, the video signals were sampled at 13.5MHz). The real-time video memory is arranged as 8 frames of 512 X 640 pixels.

The following programs are described :

Operation	Main Program	Subroutine
1-D Filtering:	UNIT	PHASE
2-D Filtering:	STAG2	PHAG2
	STAG3	PHAG3

Picture and file manipulation: FQ
COPY
AMP
SWOPF
BORD

The filter programs are described in order of their complexity.

A1.1 UNIT and PHASE

The main program, UNIT, simulates the operation of the basic one-dimensional lowpass/highpass/allpass unit shown in fig.30a. Subroutine PHASE performs the allpass FIR filtering operation associated with the phase-shifter.

The program asks for the horizontal start address of the video frame to be processed (fig. 4.2) and the horizontal start address for the output (processed) frame (these may be made equal). The program then asks for the five phase-shifter coefficients. (The coefficients are normalised in software). The coefficients are entered in the following format:

h(1)
h(3)
h(5)
h(7)
h(9)

The program will then ask for the mode of operation : lowpass, highpass or allpass.

PHASE

Subroutine PHASE simulates the half-band FIR phase-shifter in both the decimator and interpolation sections. The boundary conditions required for the FIR filtering operation default to black level.

A1.2 STAG2 and PHAG2

The main program (STAG2) simulates the two-dimensional filter structure shown in fig.4.8b. Subroutine PHAG2 performs the FIR filtering operation associated with the phase-shifters. The vertical filtering is performed on each field in turn and therefore the effective input sampling rate to the vertical filter is 312cph. The operation of both programs is similar to that of programs UNIT and PHASE.

STAG2

The program again asks for the input and output start addresses for the real-time memory (again these may be made equal) and eight phase-shifter coefficients in the form :

h(1)

h(3)

h(5)

h(7)

h(9)

h(11)

h(13)

h(15)

The program will then ask for the passband quadrants A, B, C, and D (see fig.4.8b). Enter 0 for stopband and 1 for passband. eg:

1

1

0

1

A frame store is required between the vertical decimator unit and the two horizontal decimator units (fig.4.8a). The computer will not allow an array large enough to contain a complete frame (512 X 640 pixels), and therefore the values must be stored in the real-time video memory. The lowpass and highpass results may take negative values. However the real-time video memory is only 8 bits deep (0 to +255) and therefore the values must be scaled before being written to the memory. The scaled values are checked to ensure they lie in the range 0 to +255, and are then written to the memory using system subroutine FWRTE2.

PHAG2

User subroutine PHAG2 simulates the FIR phase-shifter in the three decimator units and in the three interpolator units. Again the boundary conditions required by the FIR filtering operation default to black level.

A1.3 STAG3 and PHAG3

The main program, STAG3, simulates the same two-dimensional filter structure as STAG2. However the vertical phase-shifter acts on a complete frame (625 lines) of video information in one process, and therefore the spectrum of the filter changes to that shown in fig.4.8c. (The effective vertical sampling rate at the input to the filter has therefore been increased from 312cph to 625cph).

The software for STAG3 and PHAG3 is very similar to that for STAG2 and PHAG2.

A1.4 FQ

This program down-samples one frame of video in a field-quincunx pattern (fig.4.16) ie. alternate pixels are set to black level. The position of the black level pixels is shifted by one pixel position in the second field to obtain the quincunx pattern.

A1.5 COPY

This program copies a frame of video from one horizontal start address to another.

A1.6 AMP

This program amplifies the video signal associated with a specified frame held in the video memory. The black level pedestal is removed before amplifying the video signal, and restored after the signal has been processed. Under or overflow is corrected. The input and output start addresses are the same.

A1.7 SWOPF

This program swops the contents of the two fields of video in the specified frame (ie. lines 1 and 313 are swopped, 2 and 314 etc.). The input and output horizontal start addresses are the same.

A1.8 BORD

This program places a two-pixel-wide black border at the left- and right-hand sides of the specified frame. This program was used to suppress the overshoots which occurred at the output of the simulator. The input and output horizontal start addresses are the same.