

TV & Video Engineer's Reference Book

Edited by

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With specialist contributors

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Television Standards and Broadcasting Spectrum

Every colour television channel consists of three modulated carriers:

- The *vision* information, derived from a camera or other signal source, is used to amplitude modulate a carrier with the electrical equivalents of the basic 'black and white' variations that are encountered during transmission of the scene.
- A subcarrier, situated within the bandwidth of the vision modulated carrier, is itself modulated with information related to the *colour* information in the scene.
- A separate adjacent carrier is modulated with the *sound* information contained in the scene.

The eye, as a visual communication system, 'sees' a large amount of detail simultaneously, by virtue of the fact that it has several million communication channels operating in parallel at any instant. The electrical signals that are generated by the millions of sensors in the eye are partly processed in the retina at the back of the eye, and further processed in the brain to provide the familiar human experience of normal vision. The mass of detail forming the visual scene consists of variations in light and shade, colour and, because we have two eyes, perspective.

Picture transmission, using electronic means to convey information of a scene, cannot be carried out as a simultaneous process embracing the total field of view. Any telecommunication system can process only a single item of information at a time, and hence the data relating to any visual scene must be analysed in such a way that the complete scene can be transmitted as separate items of electrical information. At the receiver, the individual bits of information are recovered and processed for display.

1.1 Scanning and aspect ratio

The visual scene is explored by examining the small areas of detail that are contained in it, a process known as *scanning*. When we read the page of a book, our eyes scan it line by line to extract the total visual information. Electronic scanning carries out a similar line by line scan process, the detail encountered being translated into voltage variations that can be used to modulate a radio transmitter. At the receiver, the received

signals are demodulated and used to vary the beam current(s) of a display tube, the beam of which is sweeping in synchronism with the transmitter scanning beam.

A constraint of the electronic scanning system is the need to put a *frame* round the field of view to be transmitted. In the human seeing process the eye is quite unrestricted in its movements, and it roams freely over a very wide angular range which, with head and body movements, provides an unlimited field of view. In the electronic process, a finite limitation must be imposed by means of a frame, within which the picture can be analysed line by line.

The cinema industry has been in the business of presenting pictures as visual information for very many years, and has established a large number of basic principles. The newcomers to the art are the television engineers who did not 're-invent the wheel' but, very wisely, adopted many of the principles and standards that have evolved in film presentation. One of these concerns the shape of the frame. In the film industry, a standard rectangular shape with an *aspect ratio* (ar) of 4 (horizontal):3 (vertical) is the norm. This standard is still in general use for the main products of the film industry, despite the use of various 'wide screen' and other ratios. If a system has a standard aspect ratio at both the transmitter and the receiver, picture size is irrelevant. The relative dimensions of objects in the field will be correct.

The first television engineers concerned with the need to establish standards had no reason to depart from the 4:3 ar, particularly as it was realized that film would constitute a large proportion of programme material. These engineers were the team that created the standards for the world's first regular broadcast service of television programmes in 1936. This ratio has been adopted by all the systems that followed, and only recently has a change been considered in the 4:3 ratio.

Earlier experimental systems by Baird scanned the frame vertically with an aspect ratio of 1:2.

1.2 Still and moving pictures

A variety of picture transmission methods have been in use for many years. Still pictures have been communicated over telecommunication links by means of a *facsimile* (fax) system. The picture to be transmitted is wrapped round a drum, and

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scanned line by line as the drum is rotated and advanced with each turn. At the receiver, a photo-sensitive paper, wrapped round a similar drum, is synchronously rotated past a light beam which is modulated by the received signals. A high-quality 250 × 200 mm picture can be transmitted over a voice communication circuit in about 12 minutes. (Some modern fax scanners use linear flat scanning, similar to an office duplicator.)

The difference between scanning and transmission of a still or moving picture is one of time. Transmission of a still picture can take as long as we wish, but a moving field of view must be totally scanned in a time that is very short compared with the time being taken by any movements in the field of view. In other words, complete scanning of the moving picture must be so fast that we are concerned with what is, virtually, a still picture.

One of the properties exhibited by the human eye is *persistence of vision*. When the image of a still picture is impressed on the eye, removal of the visual stimulus does not result in an immediate cessation of the signals passed to the brain. An exponential lag takes place with a relatively long time for a total decay of the image. The cinema exploits this effect by presenting to the eye a succession of still pictures (or *frames*) one after the other, each frame differing from the previous one only by the change in position of any moving objects in the field of view. The presentation of one frame after another must not allow time for the image decay to become obvious and, provided that the presentation is sufficiently rapid and not too bright, an illusion of continuous movement is maintained.

The still pictures are projected in succession onto a screen. A frame is drawn into position with light cut off by means of a rotating shutter. As the shutter opens, the frame is stationary and the projected image illuminates the screen. The shutter cuts off the light, the next frame is drawn into position, the light is re-exposed through the frame, and so on in a continuous sequence. A great deal of early work with film showed that, for most people, a projection rate as low as 10–12 frames/second is adequate to present a complete illusion of movement.

However, at this projection rate another property of the eye becomes significant. The eye is extremely sensitive to the interruption of light at this rate, and the viewer would be very aware of *flicker*. As a consequence, the standard rate adopted for projection was 16 frames/second, well above that necessary for the presentation of continuous movement and, with the low-level illuminants of those days, adequate to minimize any awareness of flicker.

The eye sensitivity to flicker is a function of picture brightness and interruption rate, the rate needing to be higher if brightness is increased. With improvements in projector lamps over the years, flicker became a problem. Raising the projection rate would reduce flicker, but would result in a larger quantity of expensive film being required. An ingenious solution of this problem became a standard feature of all film projection systems. The film is drawn into position with the light cut off by the shutter, as previously explained. The shutter exposes the light through the film, cuts it off, re-exposes the light through the same frame, cuts it off and only then is the next frame drawn into position for the next double exposure. The light is projected and cut off twice during each frame. For a picture projection rate of 16 frames/second, the interruption frequency is raised to 32 per second, and the visibility of flicker is very much reduced without doubling the length of film.

As time passed, better illuminants came into use, and flicker reappeared. This problem was solved along with another concerned with the sound track that films now required. Film was not passing through the projector system fast enough for good sound quality. The standard was changed to that in use today. The frame rate was raised from 16 to 24 frames/second.

This raised the interruption frequency to 48, and increased the sound track length by 50 per cent.

1.3 Television picture frequency

The engineers who had the task of establishing standards for the first broadcast system (system A, see *Table 1.1*) adopted the aspect ratio of 4:3, but were concerned by the film projection rate of 24 frames/second. It was feared that, with a 50 Hz power supply frequency, any residual 50 Hz or 100 Hz power supply ripple in the receiver might modulate the beam current of the display tube with a sub-harmonic at 25 Hz which would produce a visible 'bar' across the picture. If film was being transmitted at the film standard rate of 24 frames/second, the difference frequency of 1 Hz would result in the bar sweeping down the picture once per second.

It was decided that the picture rate would be 25 pictures/second instead of the film rate of 24. It was considered that the effect on sound would not be serious. It was further considered that, in the event of interference from the power supply, a stationary bar across the picture would be less offensive than a bar sweeping down the picture at the 'beat' frequency of 1 per second.

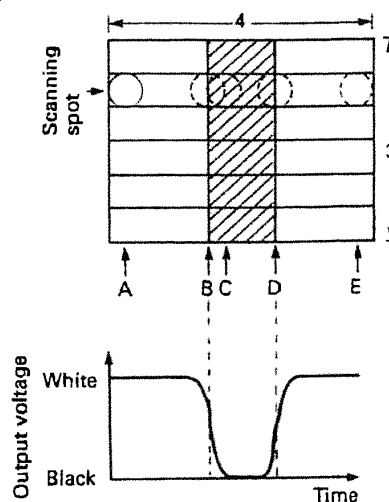


Figure 1.1 A six-line scan of a 4:3 field, showing aperture distortion

Figure 1.1 shows a simple six-line picture consisting of a black bar on a white background and the voltage output from the scanning system during a one-line scan. The scanning spot has a diameter equal to the width of one line, and sweeps across the picture from right to left along line 1, returns to its starting point displaced vertically by one line, sweeps line 2, and so on down the field of view to line 6. It is then returned to the top of the picture for the second picture scan, and so on. The voltage output may be as shown with maximum voltage indicating peak white, and minimum voltage corresponding to black, termed *positive modulation*. In a system, the polarity may be inverted, so that minimum voltage equals white and maximum is black. This would be *negative modulation*, as exemplified in the current UK system I (see *Table 1.1*). This method of scanning is termed *sequential scanning*, and results in a very obvious flicker because the field rate and picture frequency are the same, i.e. 25 per second.

A scanning system was standardized which provided a similar effect to the double-shuttering used in film projection. Instead

of scanning the lines in sequence, the picture field is scanned by one field using lines 1, 3 and 5, and a second scan fills the gaps by re-scanning the field using lines 2, 4 and 6. This is *interlaced scanning*, and constitutes two sweeps of each field for a complete picture field. This has the same effect as double-shuttering of film, and raises the flicker frequency to 50 per second.

The interruption rate imposes a limitation on the brightness level at which the display tube can be operated before flicker becomes visible. The relationship between flicker and brightness is expressed by the Ferry-Porter law:

$$f_c = F + 12.6 \log_{10} B$$

where f_c is the critical frequency below which flicker is observed, F is a constant related to viewing conditions, and B is the luminance of picture highlights.

Tests on the viewing conditions of television pictures have suggested a value of about 37 for F and, with $f_c = 50$ (as in all European and some other systems), a picture highlight value of about 10 foot-lamberts is obtained.

Television standards in the USA adopted the same general principles. The picture rate was related to a power supply frequency of 60 Hz, resulting in a picture rate of 30 per second and a light-interruption rate of 60 per second. This increase of interruption frequency, compared with the UK rate, results in a permissible increase in highlight value by 6.8 times.

The six-line system illustrated in Figure 1.1 would have very poor picture quality. The picture has a sharp transition at the edges of the black bar, but the voltage output does not change instantly from the white value to the black value. At position A (Figure 1.1) the scanning spot 'sees' peak white. As the spot reaches the bar at B it 'sees' half white and half black, the output being a half of the peak value, as shown. The output only reaches the value due to black at position C. At D, the half value is derived as shown, and the remainder of the scan gives a white output. The resulting effect is termed *aperture distortion*; it prevents any small detail in the picture being reproduced accurately.

In any practical television system, the picture quality will depend on the ability of the system to reproduce at the display tube all the sharp edges and fine detail. This requires the scanning spot size to be reduced, with a consequent increase in the number of lines necessary for a complete scan of the field of view. It is shown in section 1.5 that the channel bandwidth is determined by the scanning spot size. The smaller the spot, the more lines that are required for a complete scan, and channel space is at a premium. This means that a 'standard' spot size has to be a compromise between the ability of the system to provide a picture quality that is acceptable, and the minimum demand for channel space. Such standards are quoted in terms of the number of lines that are required in the vertical dimension for complete scanning of the entire field of view.

1.4 The video signal

The video signal derived by the camera or other scanning device for a practical black and white system will consist of random

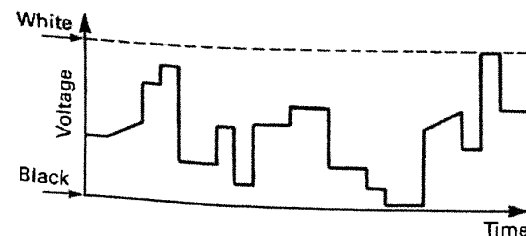


Figure 1.2 A possible video output from a camera during a one-line scan

voltages generated by the scanning of black, white and grey images during the line scan. A possible scan output is shown in Figure 1.2.

Two important conclusions arise from consideration of this type of voltage waveform:

- Voltage variations will generally consist of 'step' changes from one value to another. Smooth transitions from white to black, or from black to white, will be rare.
- AC voltage variations are extremely unlikely, and their rare appearance might arise from a scan across regular bars, such as the black and white bars on a test card.

Figure 1.3 illustrates another important feature that arises from this type of waveform. The two line scans each show the same signal voltage variation. In (a) a white bar is shown on a grey background, and in (b) the same output voltage variation shows a grey bar on a black background. The difference between the two identical signal variations is due to there being, in each, an average dc voltage component. The dc level determines picture brightness.

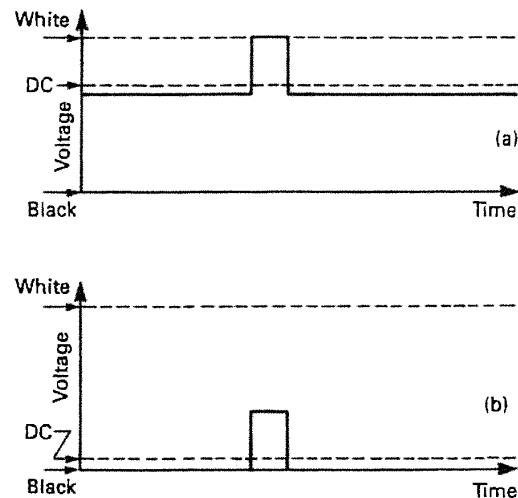


Figure 1.3 Similar video output signals are shown in (a) and (b), but with different dc levels

The video signals resulting from the scanning operation are processed and used to amplitude modulate the transmitter output. All standard broadcast transmitters use am, but fm is employed for certain links, and for satellite systems.

The principles of amplitude modulation are well known, but there are some important differences between sound and video as modulating signals. Figure 1.4(a) shows an alternating current variation that might be measured in the antenna system of a sound transmitter. Initially, the carrier is not modulated, then one cycle of an audio modulating tone is applied. The familiar features of this process are:

- The unmodulated carrier is radiated, and its mean level is constant with or without modulation being present.
- The carrier peak level is varied at the modulating frequency. The audio variation in carrier peak values during modulation is termed the *envelope*.
- An obvious limit exists in the modulating process whereby the carrier peak must not exceed twice the unmodulated level if distortion due to *clipping* is to be avoided.

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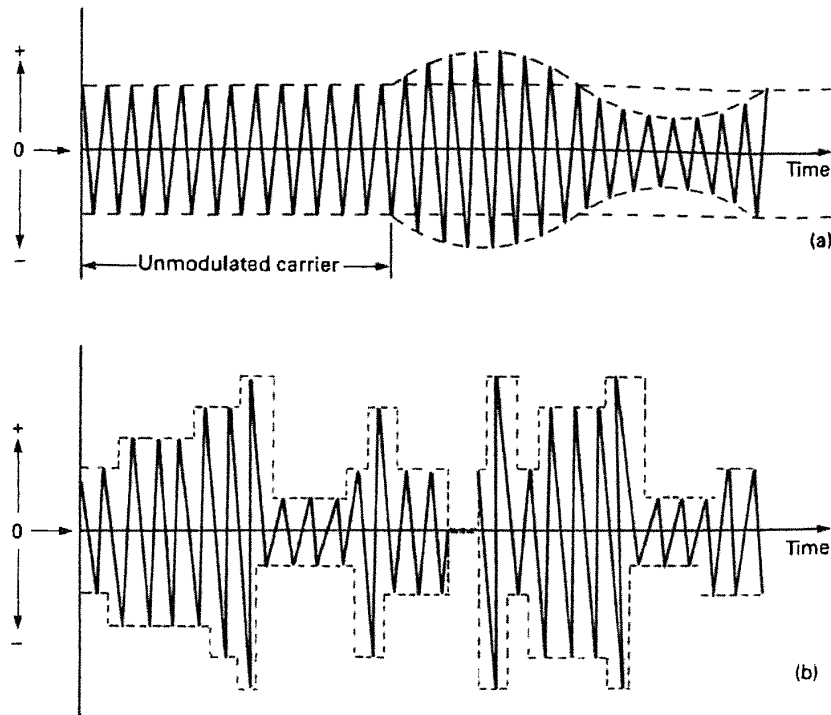


Figure 1.4 A carrier is shown amplitude modulated (a) by an audio tone and (b) by a video signal

Figure 1.4(b) shows a similar situation; it is amplitude modulation by a video signal similar to Figure 1.2. The envelope is now of a random character, and there is no mean carrier of constant level during modulation. When no modulating signal is present, no transmitter output is radiated.

height equal to the scanning spot diameter. The total number of squares in a 4:3 ar frame will be $(6 \times 6)4/3 = 48$. If we transmit 25 complete pictures per second, the squares will be scanned in $1/25$ s so that $48 \times 25 = 1200$ squares will be scanned in one second.

1.5 Channel bandwidth

If a carrier is amplitude modulated with, for example, a 1 kHz tone, three frequencies are produced: the carrier, a frequency lower than the carrier by 1 kHz (a *lower side frequency*), and a frequency higher than the carrier by 1 kHz. If the modulating signal is a band of frequencies such as voice, music or video, a band of frequencies is generated each side of the carrier, termed *sidebands*, extending on each side of the carrier frequency to a limit determined by the highest frequency in the range of modulating frequencies.

There are several ways in which the highest frequency component in a modulating video signal may be determined, and thereby the channel bandwidth. One is indicated in Figure 1.5 in which (a) is the top left-hand corner of a picture consisting of a regular pattern of alternate black and white squares. The sides of the squares are equal in length to the scanning spot diameter, and consequently the output signal generated will be a sine-wave, the frequency being the highest that will be generated at full amplitude. Any detail of smaller dimensions will not generate maximum output.

Consider now the six-line picture discussed in section 1.3, where this picture is of the pattern shown in Figure 1.5, i.e. alternate squares of black and white, each with a width and

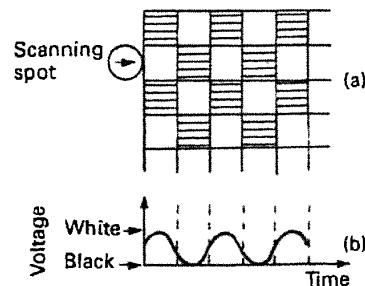


Figure 1.5 The smallest detail that can be resolved at full amplitude

The resulting ac cycle corresponds to the scan of one black and one white square. Thus the highest modulating frequency is $1200/2 = 600$ Hz, and the two sidebands would require an overall rf channel bandwidth of $2 \times 600 = 1200$ Hz.

We have seen that the picture quality would be poor, and severely lacking in the resolution of edges and fine detail. We know that, to improve resolution, we must reduce the scanning spot diameter and increase the number of scanning lines appropriately. For example, if we halved the spot diameter of our Figure 1.1 model, we would have to double the number of lines for a complete picture scan, and then the total number of squares would be $12 \times 12 \times 4/3 = 192$ and the number of squares

scanned in 1 s would be $192 \times 25 = 4800$. This would result in a channel bandwidth of 4.8 kHz, four times that of the original six-line system. The bandwidth increases as the square of the number of lines.

A standard adopted for a good quality television system has to be a compromise between the need for acceptable definition of edges and fine detail in a picture and the overall bandwidth of the channel.

There are two historical examples of 'line standards' that are worth consideration. The first UK television broadcast system developed by Baird used an aspect ratio of 1:2 and $12\frac{1}{2}$ pictures per second. Scanning was sequential and vertical. Thus, the number of squares would be $(30 \times 30) \times 2 = 1800$, and the number scanned in 1 s was $1800 \times 12.5 = 22\,500$. The highest modulating frequency was thus 11 250 Hz.

The second example was the first ever 'high definition' broadcasts that commenced in 1936 in the UK with a picture frequency of 25 per second, interlaced scanning and an aspect ratio of 4:3. The compromise on definition and bandwidth was decided on the basis of the scanning spot being of such a size that 405 lines would be required to cover the picture area, and give acceptable picture quality. The highest video frequency is $405 \times 405 \times \frac{1}{3} \times \frac{1}{2} = 2.7$ MHz. The output of detail smaller than $1/405$ of picture height would be less than maximum. Detail generating 3.0 MHz, for example, would be about -3 dB. The total video channel rf bandwidth becomes 3.4 MHz and the lowest modulating frequency is zero or dc. The total radio spectrum space occupied by this first system A (as it became known) is shown in Figure 1.6.

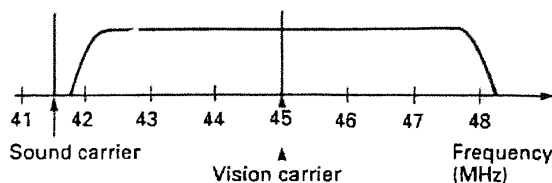


Figure 1.6 The full bandwidth of a dsb system A channel

1.6 Synchronism between scanning systems

The waveform in Figure 1.4(b) could represent the transmitter modulation during a one-line scan. At the receiver, the modulating signal is recovered and, after suitable processing, is used to modulate the beam current of the display tube, thus recreating the detail seen by the scanner during the one-line scan.

Two further important items of information are necessary to ensure that the scanning spot at the transmitter, and the beam position on the face of the display tube, occupy identical positions in their 4:3 frame. One determines the position of the scanning spot in the vertical plane, and the other ensures the correct position in the horizontal plane.

It would appear from Figure 1.4(b) that the video waveform is very complete, and there is no way in which any additional information can be provided, but a development of system A showed how it could be done, in a manner that forms part of any television standard today. Figure 1.7(a) shows how the video signal modulation can be established between the two limits: the maximum transmitter output, and the carrier level corresponding to black. This leaves a region between zero carrier output and the black level into which we can put extra information.

However, the entire transmitter/receiver system can deal with only one bit of information at any instant, and we must remove the video information while we provide any extra

information. A *blanking pulse* blanks out all video information down to the black level, at the start of the line scan. A narrow *line synchronizing pulse* is then inserted from the foot of the blanking pulse down to zero. The leading edge of this pulse is used to start the line scan on its traverse across the field from left to right. At the receiver, the leading edge of the demodulated narrow pulse is used to start the sweep of the display tube beam across the face of the tube. At the end of the line scan, the next blanking/synchronizing pulse triggers the return of the spot to the left-hand side, and the scanning cycle starts again, this time slightly displaced vertically, to trace a new line path.

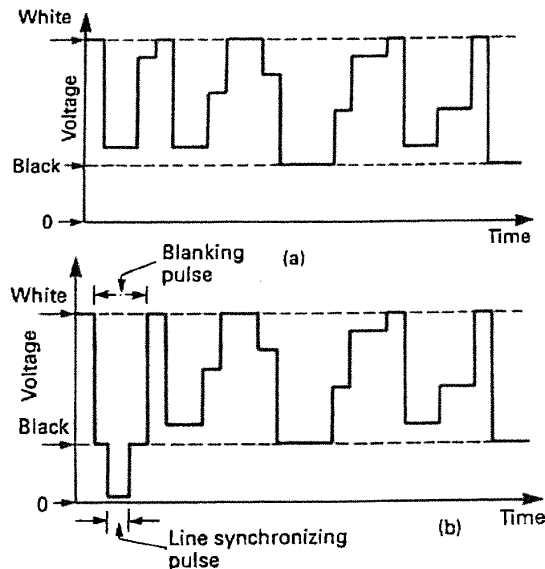


Figure 1.7 A typical camera output (a) before and (b) after the insertion of blanking and line sync pulses

The waveform of a line of the UK system I is shown in Figure 1.8. The heavy black line is the waveform of a black and white system, and the shaded areas are concerned with the colour information. The timing of all the pulses and other features is with reference to the leading edge of the narrow sync pulse. All television standards use the same type of waveform, but their timing and pulse widths may differ.

To effect vertical displacement of the scanning spot in synchronism, it is necessary to use a different type of pulse. The line sync pulses are narrow and relatively infrequent, so we can suppress video information for several lines, and transmit either a single, long pulse or a train of relatively broad pulses to effect vertical synchronism. Figure 1.9 shows the four-field sequence of field sync pulses used in the UK system I. The 'burst' sequence is necessary for the PAL colour system.

At the receiver, the two types of pulse can be separated from the composite waveform by relatively simple forms of amplitude discrimination, and they can then be separated from each other by the passive circuitry shown in Figures 1.10 and 1.11.

Figure 1.10 shows one form of a *differentiation circuit* in which the time constant CR is short compared with t . It will be recalled that synchronizing information is required from the leading edge of the line sync pulse if all the timing is to be correct, and this type of circuit provides this discrimination between leading and lagging edges.

Figure 1.11 shows an *integrating circuit*. The time constant CR is long compared with the duration of the train of pulses provided at the input. Successive pulses build up the voltage on C to the value V_p and, at the end of the pulse train, C

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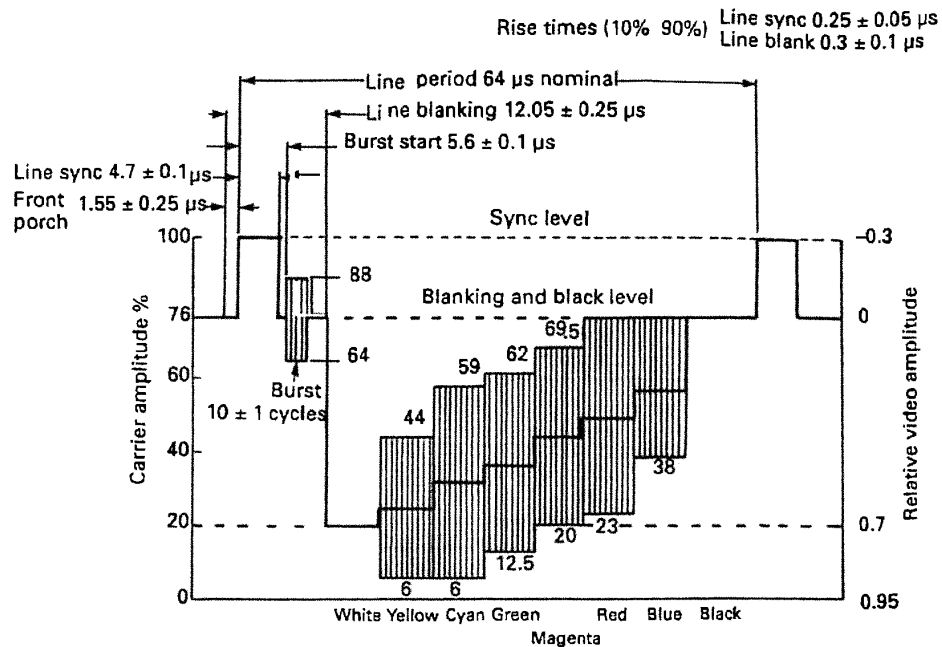


Figure 1.8 The thick line show a possible signal output resulting from a single line scan of a black and white system, the image being a grey 'staircase' from white to black. The shaded areas are concerned with the addition of colour (BBC)

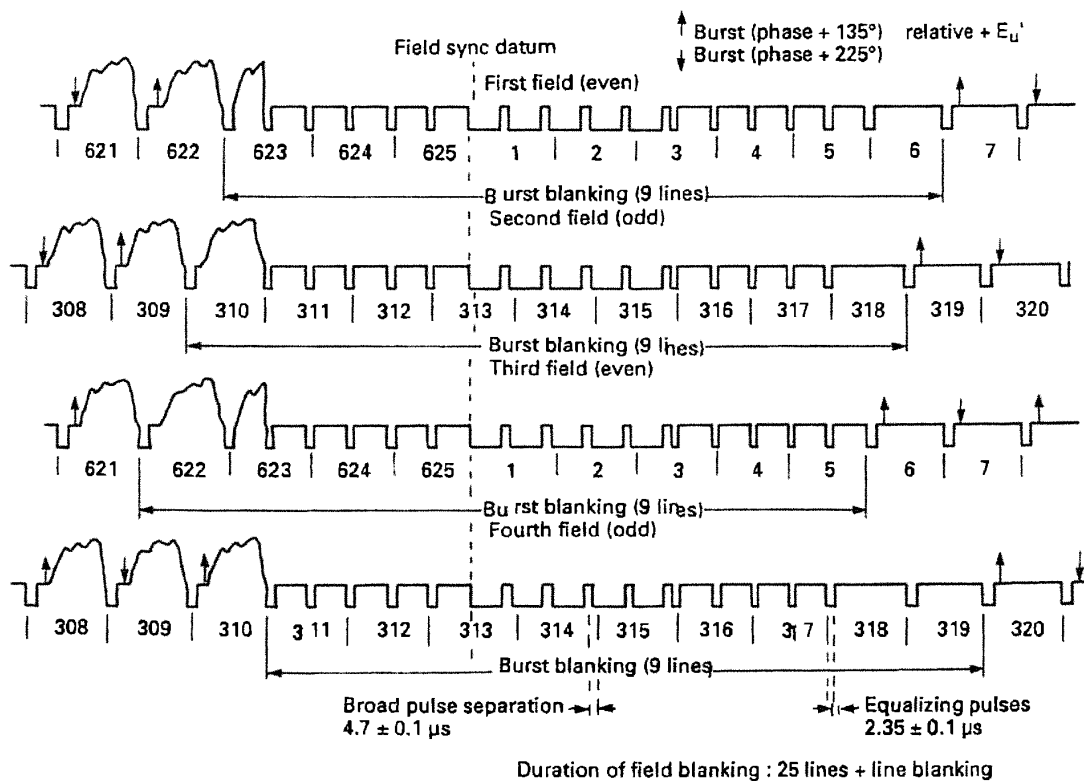


Figure 1.9 The broad pulses used for synchronism of the vertical scan during blanking of 25 lines

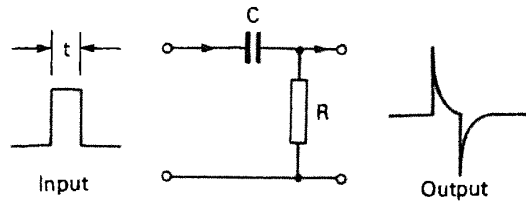


Figure 1.10 A differentiation circuit that provides an output when the input voltage changes in value

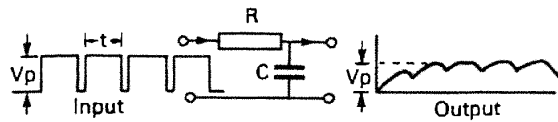


Figure 1.11 An integration system that enables a train of broad pulses to build up to a peak value

discharges. The output waveform thus constitutes a single broad pulse, which operates at a relatively slow speed to trigger the vertical sweep system.

The differentiation circuit does not distinguish between line or field pulses; it will generate 'spike' pulse output from the edges of any pulse fed into it. The integration circuit is the one that *does* discriminate between the two types of pulse. The narrow, infrequent line sync pulses will provide no output from the integrator.

1.7 Porches

Figure 1.8 shows two other features that are concerned with the line sync pulse region. Note that the pulse is not in the centre of the blanking interval.

In any system that contains inductance, capacitance and resistance, voltage or current changes cannot take place instantaneously. A video signal change from black to peak white, or white to black, takes time for completion, requiring possibly a capacitor to charge or discharge. A black/white edge results in a voltage change as shown in Figure 1.12 (not to be confused with the aperture distortion shown in Figure 1.1). Ahead of the pulse in Figure 1.8 is a narrow plateau, the *front porch*. Its purpose is to allow the video signal of the previous line (which might have been at peak white, for example) to reach black level before the sync pulse starts.

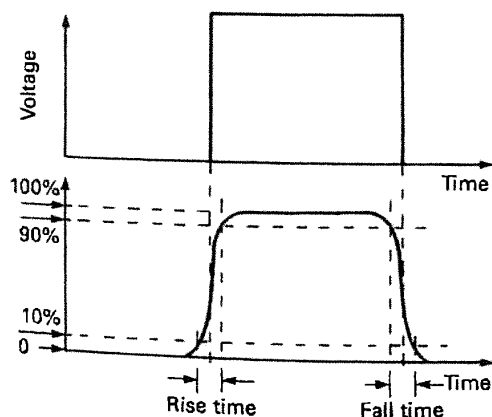


Figure 1.12 Voltage and current values cannot change instantly. Rise or fall times are determined by the values at 10 per cent and 90 per cent of peak values

Behind the sync pulse is another plateau, the *back porch*. The original purpose of the back porch, when first used in system A, was to make sure that there was plenty of time for the receiver line scan circuitry to effect complete retrace of the line scan, and for the beam to be in its correct position for commencement of the next line scan. By comparison with current technology, receiver scan circuits in 1936 were crude, ponderous and extravagant of power, and they required a lot of time for retrace. Today, the back porch provides plenty of time for retrace and it also provides the space necessary for colour information to be transmitted and extracted at the receiver.

1.8 DSB, ssb, asb and vsb

The UK system A operated from 1936 to 1939, and then had to close down due to the outbreak of World War II. The USA standardized its television system and immediately found a serious problem. The double sideband (dsb) system was very extravagant of channel space and, as many channels were required, a new system of modulation was devised. It was known as a *vestigial sideband* (vsb) system, sometimes termed an *asymmetric sideband* (asb) system, and is now used throughout the world because it saves a considerable amount of channel space.

To appreciate the operation of vsb, it is necessary to digress into some important differences between sound and video amplitude modulation. Single sideband (ssb) am systems have been used for radio transmission of speech and music for at least 60 years. Why cannot ssb be used for video modulation and thus save half the dsb bandwidth?

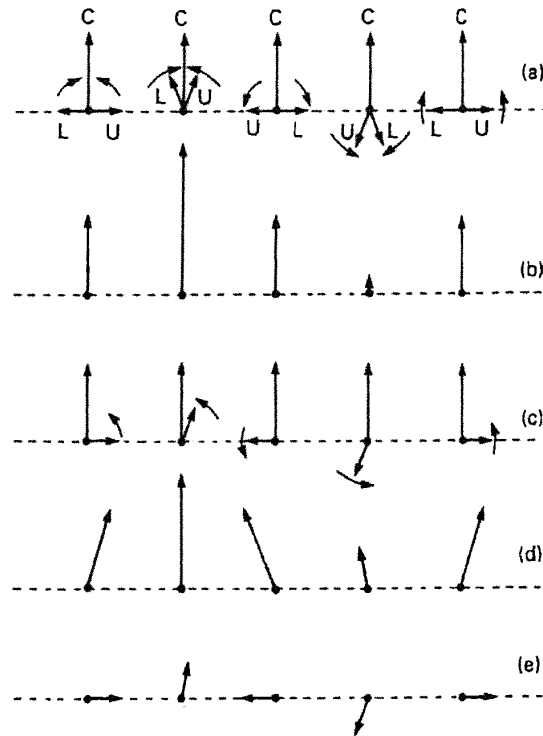


Figure 1.13 The vector relationships of dsb (a) and (b) ssb (c) and (d), and a single side frequency (e)

There are two sources of distortion in a dsb speech or music system. An amplitude constraint exists whereby any attempt at

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over-modulation results in severe waveform distortion, with unacceptable audible quality. The second effect occurs in single-channel sound reproduction where the quality of the reproduced sound is unaffected by changes of phase response. Many do not accept that our normal binaural experience distorts. Nevertheless, simply turning one's head, for example, results in a severe phase change of, say, a 5 kHz tone, where about 4 cm movement represents about 180° . We live with this effect and do not notice it, even when we use it for directional information.

Figure 1.13(a) shows the vectors relating to the carrier side frequencies during one cycle of a modulating tone of dsb am at maximum modulation; (b) shows the resultant vector addition of the vectors in (a), causing the rise and fall of carrier amplitude, and the phase of the resultant vector, which remains that of the carrier at all times during the modulating cycle.

Figure 1.13(c) is as (a), but with the lower side frequency suppressed, and (d) is the new resultant. Two features can be seen:

- The modulation, considered in terms of carrier amplitude, is halved.
- The angle of the resultant vector is now swinging with respect to the carrier vector.

We have produced phase modulation.

To determine the requirements of a video signal used for amplitude modulation, consider the simple video modulating signal shown in Figure 1.14(a). Such a pulse can be analysed into a number of discrete harmonic components, which extend out to infinity with a descending order of amplitude. The square waveform of (a) could be synthesized by addition of the frequency components of (b) at the correct amplitudes and phases. In a television system, we require the pulse signal to progress through the system in such a manner that the shape of the waveform that finally modulates the display tube beam current is a faithful copy of the signal derived by the scanning process at the transmitter.

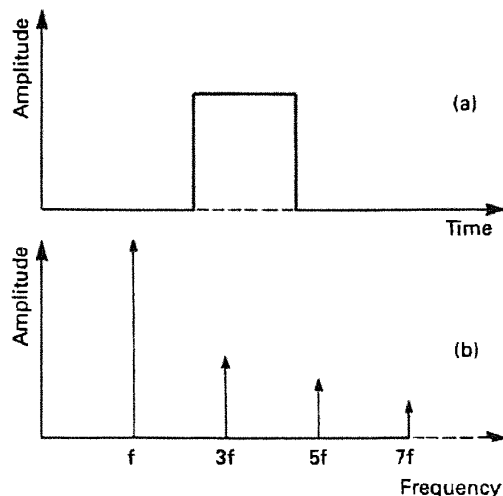


Figure 1.14 A theoretical rectangular pulse contains a fundamental frequency f , and all the odd harmonic frequencies. The practical pulse loses some of the higher harmonics due to the finite bandwidth of the system, and has sloping sides with rounded corners

A pulse can experience degradation of its waveform in a number of ways as it goes through the system:

- 1 The bandwidth of the system is finite, and therefore some of

the higher frequency harmonics will not be present.

- 2 An inadequate hf response at some point in the signal path can change the pulse shape by, for example, rounding the corners and sloping the sides of the pulse.

- 3 Inadequate lf response can produce a 'tilt' at the top of a pulse. The dc voltage may not hold up for the duration of the pulse.

- 4 The various discrete frequency components that constitute the signal waveform may experience differing transit times as they travel through the system. Some may go through faster or slower than others, distorting the wave shapes in varying degrees at the point where they are intended to arrive together, i.e. at the point where the display tube beam current is modulated.

Of the above, 1 is determined by a definition compromise as discussed in section 1.5, and 2 and 3 are concerned with circuit behaviour, and any deficiencies can be resolved. The timing effect of 4 is much more important. The phase response of the system, clearly, determines the preservation of pulse shape. Each of the signal frequency components must travel through the system in the same time, although the actual time of transit is, within reason, of no importance.

If a fundamental frequency f passes through the system in time t , this time can be interpreted as a phase change ϕ . The frequency $2f$ will go through twice this phase angle in the same time. $3f$ will shift three times the fundamental phase angle, and so on for each harmonic component. If one of the harmonic components goes through too great a phase shift in the time t , it means that it is travelling too fast and will arrive early at the tube beam current control. It is seen that the phase response of the system should be such that the angle ϕ must be proportional to frequency. Any one of the lines shown in Figure 1.15 would indicate a satisfactory phase response. The linear relationship is essential, and the only significance of the differing slopes is that they relate to different transit times. (The horizontal line would indicate zero time which is, of course, not possible.)

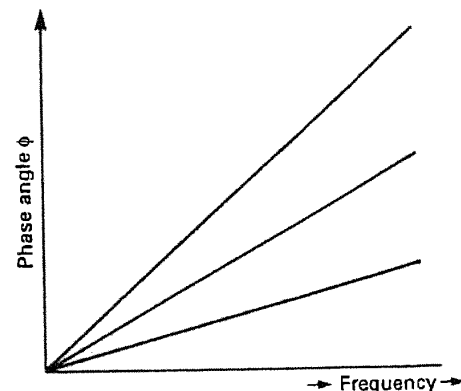


Figure 1.15 A video system phase response must be proportional to frequency

The main differences between an audio channel and video signal processing are now more clearly seen. Audio signal channels can tolerate a high degree of phase distortion, but amplitude distortion must be kept to very low limits. Video signal processing must ensure that phase distortion is minimized, but overload effects are far less serious. The peak value of Figure 1.14 may represent peak white, for example, and any increase above this value would not be visually of much significance, although it may drive a transmitter into an

overload condition. Alternatively, the peak value may correspond to a black signal, in which case 'blacker than black' is of no visual consequence.

The phase distortion produced by ssb (Figure 1.13(c) and (d)) may be acceptable for audio, but is quite unacceptable for video. However, it is possible to use partial suppression of one sideband (vsb) for a television system and save considerable channel space. Whereas phase distortion can be quite unacceptable where large picture areas are involved, phase errors become difficult to see on small detail in the picture. Thus, a practical system must have a very low phase distortion at low video frequencies, while the high frequencies generated by the fine detail in the picture can be severely distorted, but still acceptable.

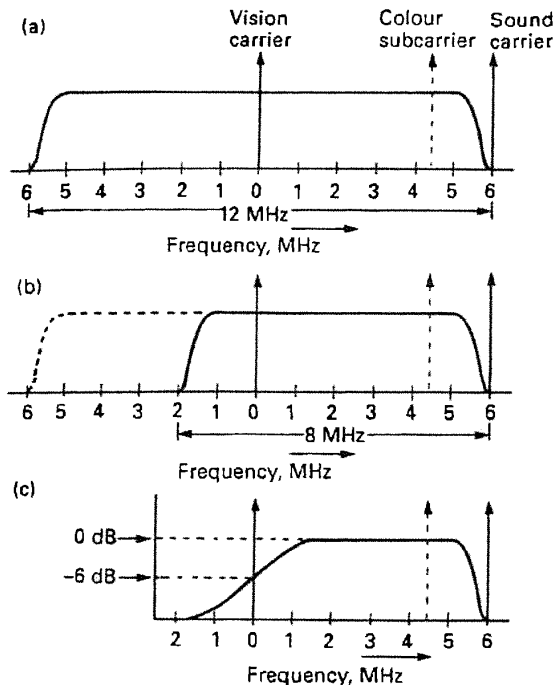


Figure 1.16 A full channel bandwidth is shown with, at (b), the radiated signal of vsb modulation, and at (c), the required receiver response

Figure 1.16 shows, at (a), the full dsb channel bandwidth that would result from video amplitude modulation, together with the sound channel, for the UK system I. The transmitted signals are shown in (b) where the lower sideband has had about 4 MHz filtered off, leaving a 'vestigial' 1.25 MHz. The 12 MHz channel has been reduced to 8 MHz.

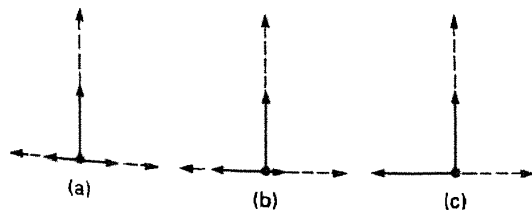


Figure 1.17 Demodulated output from the vsb input to a receiver. See description in text

For vsb to function correctly, it is necessary for the received signals to be processed before being presented to the

demodulator. The pass-band of the receiver if channel must be shaped as shown in Figure 1.16(c). Consider, for example, a 50 Hz modulating frequency, and its position in Figure 1.16. Such a frequency would result from the scan of a large area, indeed a complete field, and the resulting side frequencies would be very close to the carrier. The signals presented to the demodulator by the receiver if amplifier are shown as vectors in Figure 1.17(a). The carrier and side frequencies are cut by 6 dB, but no phase distortion would result because the signal is a full double sideband am signal at 100 per cent modulation.

Consider next a modulating frequency resulting from the scan of a smaller detail, producing a frequency around, say, 500 kHz. Figure 1.17(b) shows the vector relationship between side frequencies and carrier. The overall length of the upper plus lower side vectors add to the same value as in (a), showing that the modulation factor has not changed. However, one vector is longer than the other, and some phase distortion is beginning to appear in the resultant addition of the three vectors. At frequencies of about 1.5 MHz and above, the vector situation is as shown in (c). The single side frequency resulting from the scan of a small detail has an amplitude equal to the attenuated carrier and, again, the level of modulation is unchanged. Phase distortion is present, but the detail is small and the errors become more difficult to see as the modulating frequency increases.

The US problem of minimizing channel space was eased considerably by the use of vsb, and all standard systems now use vsb. All UK system A stations that followed Channel I used vsb and, when the first station moved from its original site to Crystal Palace, the opportunity was taken to bring it into line and change it from dsb to vsb. The vestigial band was about 1 MHz, and the overall channel width was 5 MHz.

1.9 National standards

Many of the world's standards use 625 lines, the exceptions being the 525-line system M, used by North and South America, Japan and a few other countries, and the French system E on 619 lines. System E was developed before World War II in an attempt to provide a better picture quality than the 405-line standard then prevailing. This virtual doubling of the line structure does, indeed, provide an excellent picture quality, although at the expense of channel space. Consideration is currently being given to the adoption of new line standards that will approximately double the number of lines.

The almost universal adoption of a 625-line standard resulted from the first international conference on television standards after World War II. All countries planned to start a television service, and it was hoped that the 625-line standard would permit international links between European and other countries that used 625 lines. Unfortunately, widespread adoption of 625 lines did not result in a universal standard. There are many 625-line systems, but few are compatible. Some have positive modulation, others negative. Some have am sound, and others fm sound. There are further differences in the colour.

The US system M was the first practical broadcast colour system. A later European conference attempted to standardize a European colour system. It was considered that the NTSC system M could be improved, and several proposals were made for systems based on the NTSC principle that used a subcarrier for colour information. Germany proposed a modified NTSC system known as PAL (*Phase Alternation Line*). The French proposal was for the subcarrier to be frequency-modulated, the system being termed SECAM (*Séquentiel à Mémoire*). Other variants of the NTSC system were considered for a European standard, but the final outcome was that PAL was preferred

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Table 1.1 Standard television systems

	A	B	C	D(K)	E	G	H	I	L	M	N
Lines per picture	405	625	625	625	819	625	625	625	625	525	625
Field frequency, Hz	50	50	50	50	50	50	50	50	50	60	50
Line frequency, kHz	10.125	15.625	15.625	15.625	20.475	15.625	15.625	15.625	15.625	15.734	15.625
Video bandwidth, MHz	3	5	5	6	10	5	5	5.5	6	4.2	4.2
Channel bandwidth, MHz	5	7	7	8	14	8	8	8	8	6	6
Sound/vision carrier spacing, MHz	3.5	5.5	5.5	6.5	11.15	5.5	5.5	6	6.5	4.5	4.5
Vestigial sideband width, MHz	0.75	0.75	0.75	0.75	2	0.75	1.25	1.25	1.25	0.75	0.75
Vision modulation polarity	+ ve	- ve	+ ve	- ve	+ ve	- ve	- ve	- ve	+ ve	- ve	- ve
Sound modulation	am	fm	am	fm	am	fm	fm	fm	am	fm	fm
Deviation, kHz		50		50		50	50	50		25	25
Pre-emphasis, μ s		50		50		50	50	50		75	75

and adopted by most countries, while France and Russia decided to use SECAM. (They actually have different versions of SECAM.)

A common feature of all colour systems is that they use a subcarrier for colour information, and transmit it in the form of a 'burst' of a few cycles for use by the receiver as a reference. The position for the burst is on the back porch, as shown in Figure 1.8.

Table 1.1 gives details of most of the standards in use throughout the world. (System A has been included, although it is no longer in use in the UK.) These are not the only differences between systems, but the radio frequencies on which they operate also differ between countries. The frequency allocations on which radio services operate are decided by the International Telecommunications Union (ITU). The world is divided by pole-to-pole boundaries into three regions:

Region 1 includes Europe, Africa and Russia. The eastern boundary includes the whole of Russia and Mongolia.

Region 2 includes North and South America, Greenland and Alaska.

Region 3 includes Australia, New Zealand, India and Pakistan, China and Japan.

This partition is intended to ensure that the various radio services should operate with minimum interference to each other and other services. However, this does not ensure that all television services will operate on the same frequency bands. Considerable differences exist, not only between regions, but within regions.

For example, Table 1.2 shows the UK and European allocations for television services, but Africa and Scandinavia, also in Region 1, differ. South Africa has two bands, one on vhf 175–255 MHz and the other a uhf band on 471–632 MHz. The Danish bands are 55–68 MHz, 175–216 MHz and 615–856 MHz. Both countries use PAL. South Africa uses system I, as in the UK, and Denmark uses PAL on systems B and G which differ only in channel bandwidth.

Many similar examples can be quoted. A notable one is that Japan uses NTSC system M, but a receiver manufactured by the Japanese for home use could not be used in the USA because the two countries are in different regions and their radio frequencies are different.

Further examination of Table 1.1 shows several very similar standards. Many have derived from basic systems, and differences are small. Systems D and K, for example, are the same for the listed parameters, but the D version uses PAL and the K version uses SECAM. The K system, in turn, has spawned the K' system, which is similar to the K system but with a wider vestigial sideband.

1.10 Bands and channels

The radio frequency spectrum is divided up into bands as follows:

vlf	<30 kHz
lf	30–300 kHz
mf	300–3000 kHz
hf	3–30 MHz
vhf	30–300 MHz
uhf	300–3000 MHz
shf	3–30 GHz
ehf	30–300 GHz

The band classifications are international, and broadcasting takes place in some of them. Medium-wave broadcasting, for example, occurs in the mf band. The '3–30' range of each band is not so arbitrary as it may appear (see section 12.2.1). Each band is characterized by its own propagation and antenna features.

Table 1.2 European bands and designations

Band	European band	Frequency range
vhf	I	41–68 MHz
vhf	II	88–108 MHz (fm sound)
vhf	III	174–216 MHz
uhf	IV	470–582 MHz
uhf	V	614–854 MHz
shf		11.7–12.5 GHz

Notes:

1. The UK used bands I and III for a television service, on standard A. This has now been discontinued, but other parts of Europe still use these bands for television.
2. Band II is not used for television, but for fm sound only.
3. The shf band has been allocated for satellite television services.

The sections of the spectrum that are allocated to television are in the vhf and uhf bands. There are three world zones with different frequency allocations for broadcasting, and Table 1.2 shows the European bands and their designations.

The UK television service functions in Bands IV and V only. Channels 21–34 are in Band IV, and channels 39–68 are in Band V. The various standards in use have different channel bandwidths so, for example, a UK channel 40 does not necessarily occupy the same frequency space as a European channel 40.

Table 1.3 UK shf frequencies

SHF channel	Frequency GHz	
4	11.78502	
8	11.86174	
12	11.93846	all left-hand circular polarization
16	12.01518	
20	12.09190	

The actual frequencies in the shf band allocated to television broadcasting in the UK are shown in Table 1.3. There are 40 channels, each 20 MHz wide. The precise form of the transmission has yet to be decided.

1.11 Adding colour to a monochrome system

Colour television was demonstrated in 1928 by Baird, using red, green and blue filters before each field in a sequential manner, with appropriate synchronized filters before a modulated white light source at the receiver. Various unsatisfactory features appeared, such as *colour break-up* (a fast-moving object in the field was trailed by a colour), and flicker, requiring an increase in field rate.

A more sophisticated version of a mechanical sequential colour field system was proposed by CBS when the US sought a standard for a colour system, and this proposal was adopted in 1951, only to be subsequently rejected when the US National Television Standards Committee (NTSC) was formed to derive a wholly electronic standard. The NTSC system was adopted by the US in 1953, and had its first broadcast in 1955.

Modern colour television has resulted from the addition of suitable techniques to existing monochrome systems. The transmission, reception and processing of black and white signals require three separate items of information: the video information and the synchronizing information for two-dimensional scanning.

It is shown in Part 3 that colour requires three further types of information. It is necessary to determine *luminance*, *hue* and *saturation* of any colour at any instant. Fortunately, existing monochrome systems are luminance-only systems, and they can supply the luminance information. This leaves two extra items of information to be transmitted and received which, when suitably processed, can determine the hue and saturation encountered during scanning.

Apart from the obvious difficulties, some severe constraints exist on the addition of colour information:

- It must not require extra channel space.
- There must be no interference between the existing luminance information and the added information.
- It must be totally compatible, i.e. a monochrome receiver must display a colour transmission in black and white, and a colour receiver must be capable of displaying a monochrome transmission in black and white.

1.11.1 Colour bandwidth

There are some aspects of human colour perception that can be used to engineering advantage. One of these concerns the bandwidth required for colour display. Our ability to perceive fine detail in colour is considerably inferior to our awareness of fine detail as a brightness or luminance variation, thus colour information does not require the same wide bandwidth that is needed by a luminance system.

The eye can resolve small variations of luminance detail over the very small angle of 0.5–1.0° (the precise figure varies with

individuals). Visual acuity for colour is far less sensitive than for luminance, and depends to some degree on the range and saturation of the colours involved. Over an orange/cyan range of colours the resolution angle becomes 1.5–3.0°, and over a green/magenta range, detail resolution requires about 4–8°. The bandwidth required for colour information can thus be reduced to between $\frac{1}{3}$ and $\frac{1}{8}$ of the width required for luminance. Whatever methods we are going to use to process colour, it is only necessary to use about 1 MHz for the relatively low-detail colour information.

1.11.2 Spectrum utilization

Let us examine more closely the distribution of energy over the band of frequencies occupied by a carrier which is modulated by a video waveform. Without any video modulation, the carrier is already modulated by a number of frequencies. The line scanning rate is at a constant frequency (15 625 lines/second in system I), and it establishes side frequencies on each side of the carrier, spaced at line frequency and from each other, as shown in Figure 1.18. Interlaced scanning modulates the carrier at 50 Hz, and the side frequency of 50 Hz and the harmonics are spaced on both sides of the carrier and the line scanning harmonics. The picture frequency is another modulating signal at 25 Hz, and its energy joins the clusters around each line frequency harmonic.

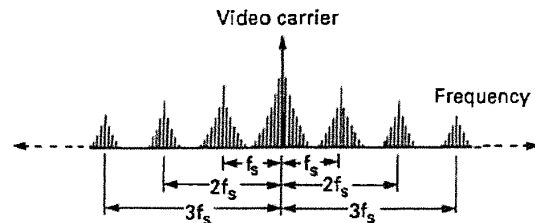


Figure 1.18 The signal components resulting from amplitude modulation. f_s is the line scanning frequency

The line frequency harmonics with their side frequencies extend out to the bandwidth limits of the video channel, and they are of a descending order of energy level as they become more remote from the carrier. Over the channel bandwidth, the energy generated across the band, without video modulation, is contained in clusters, spaced at line-scanning frequency from the carrier and from each other. Large regions of the channel space contain little or no energy.

In 1934, long before any broadcast television system existed, two mathematicians, Merz and Gray, studied the situation when video signals were included in the modulation process and concluded that the video energy simply joined the existing clusters round each line-scan harmonic, leaving the overall pattern substantially unchanged. The only differences between one picture and another, or whether one is moving or stationary, are relatively small variations in the magnitude of the energy in the clusters. The low or zero energy gaps between the line frequency harmonics remain with all types of video modulation.

The original work of Merz and Gray was directed at reducing interference between two adjacent transmitters, operating on nominally the same frequency. If their carrier frequencies were 'offset' by half the line frequency, the line harmonics of each would drop into the low-energy gaps of the other, thus minimizing mutual interference. However, in the years 1951–1953, the gaps were seen by the NTSC as the place for colour information.

A subcarrier can be inserted in the video frequency band, at a frequency about 1 MHz lower than the top of the video band,

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where the line-scan harmonics are low in amplitude. This subcarrier can then be modulated with colour information in such a manner that, when extracted and processed in the receiver, the colour display tube is driven with appropriate display information.

All existing standard colour systems are derived from the concept of the use of a subcarrier in the video band to supply colour information.

Acknowledgement

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31

DBS Receivers

A dbx receiver is capable of receiving direct-to-home signals from the broadcasting satellite in the service area. As a piece of domestic equipment, its cost must not be excessive and its performance must be adequate.

As well as the features inherent in satellite broadcasting, dbx systems are expected to improve the quality of sound and vision and to realize diversified new broadcasting services.

The receiver is an important component of the complete dbx system. Efforts have been made to realize high performance receivers at reasonable cost.

Operational dbx systems using the 12 GHz band are currently in service or in preparation, and dbx receivers are on the market in some countries. This section describes the performance of the 12 GHz band dbx receiver as a consumer product.

31.1 DBS transmission systems

In conformity with the WARC-BS plan, satellite broadcasting systems have generally adopted frequency modulation for the rf signals. In spite of the weak signals inherent in satellite broadcasting, broad-band fm makes possible the reproduction of noise suppressing picture and sound quality. The baseband signal format has been evolved to fit satellite broadcasting transmission and to improve the quality of the signals and the service flexibility.

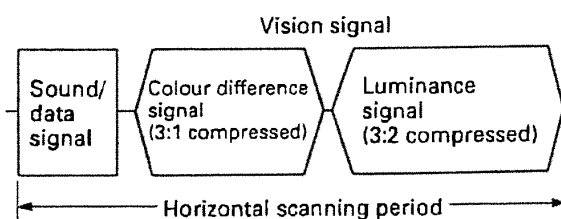
Three major dbx transmission systems are reported by the CCIR: the MAC/packet family, the B-MAC system and the digital subcarrier/NTSC system.

The MAC/packet family and the B-MAC system adopt the principle of *time-division multiplexing* (tdm), which permits an improvement in picture quality. The sound/data signals are multiplexed at rf or baseband in the line blanking interval.

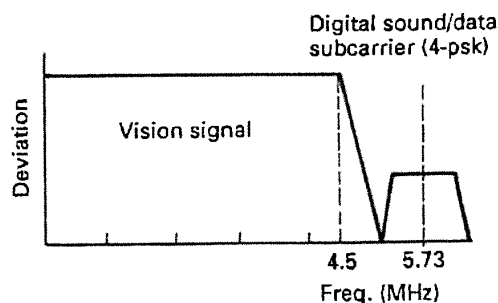
On the other hand, the digital subcarrier/NTSC system uses *frequency-division multiplexing* (fdm). The vision signals have almost the same parameters as those of a conventional M/NTSC system, for compatibility with terrestrial television. The sound/data signals are frequency-multiplexed with the vision signal.

Figure 31.1 shows the multiplexing structures of each system. These systems apply digital techniques for sound and data transmission, in order to make full use of the capacity and the flexibility provided by the channel. The sound signal comprises

the accompanying high quality sound signals and some possible additional independent sound signals. The data signal has the capability of transmitting digital data services such as teletext, facsimile and telesoftware, service identification and conditional access systems, etc.



(a)



(b)

Figure 31.1 Multiplexing structures of the dbx systems: (a) MAC/packet family and B-MAC, (b) digital subcarrier/NTSC

The signal specifications for each system are shown in Table 30.1. The principal variations between the receivers of each

31/2 DBS Receivers

Item	System	C-MAC	D2-MAC	B-MAC		Digital subcarrier/ NTSC
Country		EBU	W. Germany France	Australia	USA Canada	Japan
Nominal channel bandwidth (MHz)		27			24	27
Line/Field		625/50			525/60	
Vision coding/Modulation		MAC/fm			NTSC/fm	
Nominal sampling frequency (MHz)		13.5		14.22	14.32	—
Luminance [chrominance] compressed bandwidth (MHz)		5.6 [2.4]		5.0 [3.1]	4.2 [2.1]	4.5 [2.5/1.5]
Nominal transmitted baseband bandwidth (MHz)		8.4		7.5	6.3	6.3
Reference clock frequency (MHz)		20.25		21.328	21.477	—
Sound/data multiplexing principle		tdm* at rf	tdm* at baseband			fdm* (5.73 MHz)
Data coding/modulation		2-4 psk	Duobinary-fm	Quarternary/binary-fm		4-psk
Symbol rate (M baud)		20.25	10.125	7.11	7.16	2.048
Number of bit per symbol		1		2/1		1
Instantaneous bit rate (Mbit/s)		20.25	10.125	14.22/7.11	14.32/7.16	2.048
Mean data rate (Mbit/s)		3.08	1.54	1.59	1.60	2.048
Multiplex description		Flexible		Rigid		Flexible
Type of multiplex		Packet		Continuous		
Sound coding		14bit/sample lin* or 14/10 bit/sample nic*		Adaptive delta modulation		16 bit/sample lin* or 14/10 bit/sample nic*
Protection (Error Correction)		1 parity bit/sample or 5 bit Hamming code/sample		2.33 bits/13 bit block		BCH (63, 56), SEC, DED*
Maximum number of audio channel		8	4	6/3		4 (15 KHz), or 2 (20 KHz)

*Note: tdm: Time-division multiplexing fdm: Frequency-division multiplexing
lin: Linear coding nic: Near instantaneous companding
SEC, DED: Single error correction, Double error detection

Table 31.1 Summary of signal specifications in dbx system

system relate to the way in which the baseband signals are processed.

31.2 Configuration

A typical dbx receiver system consists of four elements: an antenna, an outdoor unit, an interconnecting cable and an indoor unit (see Figure 31.2).

The receiving antenna is installed on or adjacent to the viewer's house or building and can be pointed in the direction of the incoming signals. The outdoor unit is directly attached to the receiving antenna. The indoor unit, which contains controls such as channel selection, is located close to the conventional TV receiver.

Generally, the 12 GHz dbx receiver system uses double frequency conversion to ease the problem of selectivity, image rejection and local oscillator radiation.

31.2.1 Receiving antenna

The incoming signals from the satellite are received by the antenna. The receiving antenna consists of a reflector, primary feed, polarizer and mount with aiming mechanism. A flat antenna will consist of an antenna board, built-in outdoor unit and mount with pointing mechanism.

31.2.2 Outdoor unit

The outdoor unit amplifies and converts the incoming signal (shf; 12 GHz band) received by the antenna into the first intermediate frequency (if) signal (uhf; 1 GHz band) for transmission through a coaxial interconnecting cable to the indoor unit. It consists of a low noise amplifier (lna), a first local oscillator, a mixer, and a first if amplifier.

31.2.3 Indoor unit

The indoor unit selects the desired channel from the first if signals, converts the signal into second if, and then demodulates the fm signal to the baseband signal. It consists of a second converter, a second if amplifier, an fm demodulator, a deemphasis network, an energy dispersal signal rejection circuit, a baseband signal processor, a vhf/uhf remodulator, and a dc power supply. A baseband signal processor, a video signal processor and a sound/data signal processor are also provided.

existing TV receiver or the information processor. It may be incorporated in the TV set.

31.3 Technical features

The satellite broadcasting systems have the following inherent technical features:

- *wide-area coverage*: a wide area can be covered simultaneously

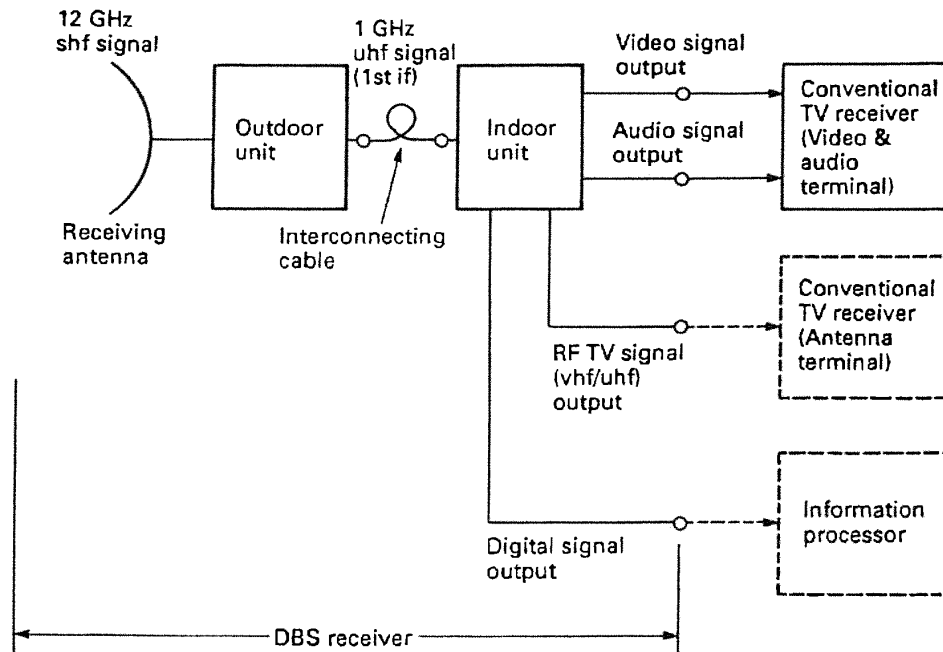


Figure 31.2 Basic configuration of dbs receiver

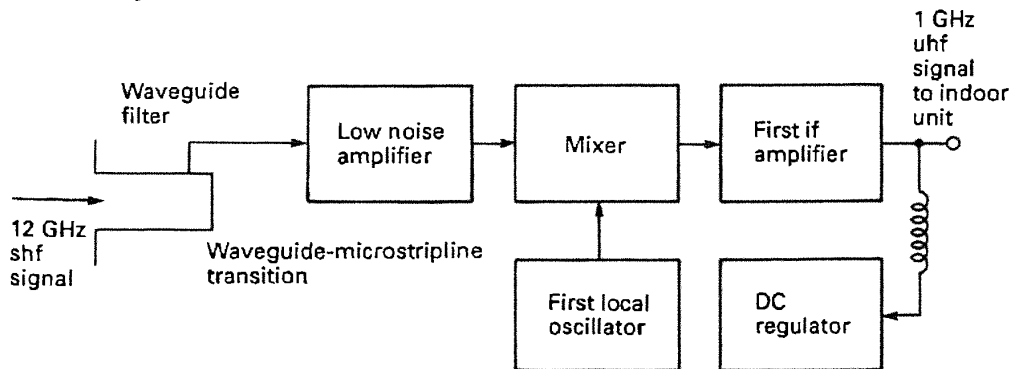


Figure 31.3 Outdoor unit

In the MAC/packet family and the B-MAC system, a MAC vision decoder is needed. For sound/data signal processing, the receiver needs a demodulator or decoder, a de-multiplexer and a digital sound decoder.

The indoor unit has various types of output interfaces, such as baseband video/audio signal outputs, rf TV signal output and digital signal output (bit-stream), etc. The indoor unit functions as an adaptor connected to the main receivers, such as the

- *high quality*: ghost free high quality picture and digitally modulated high quality sound
- *disaster immunity*: relaying is possible from anywhere via satellite
- *development potential*: introduction of new broadcasting systems is possible

In addition, the dbs receiver itself uses the following technologies:

31/4 DBS Receivers

- microwave technology for consumer use
- digital communication techniques
- digital audio and video signal processing
- data transmission and applications

Receivers are, of course, expected to comply with the technical requirements used in the planning of the WARC-BS 77 or RARC-SAT 83. For example, the recommended characteristics of home receiving equipment for satellite broadcasting in

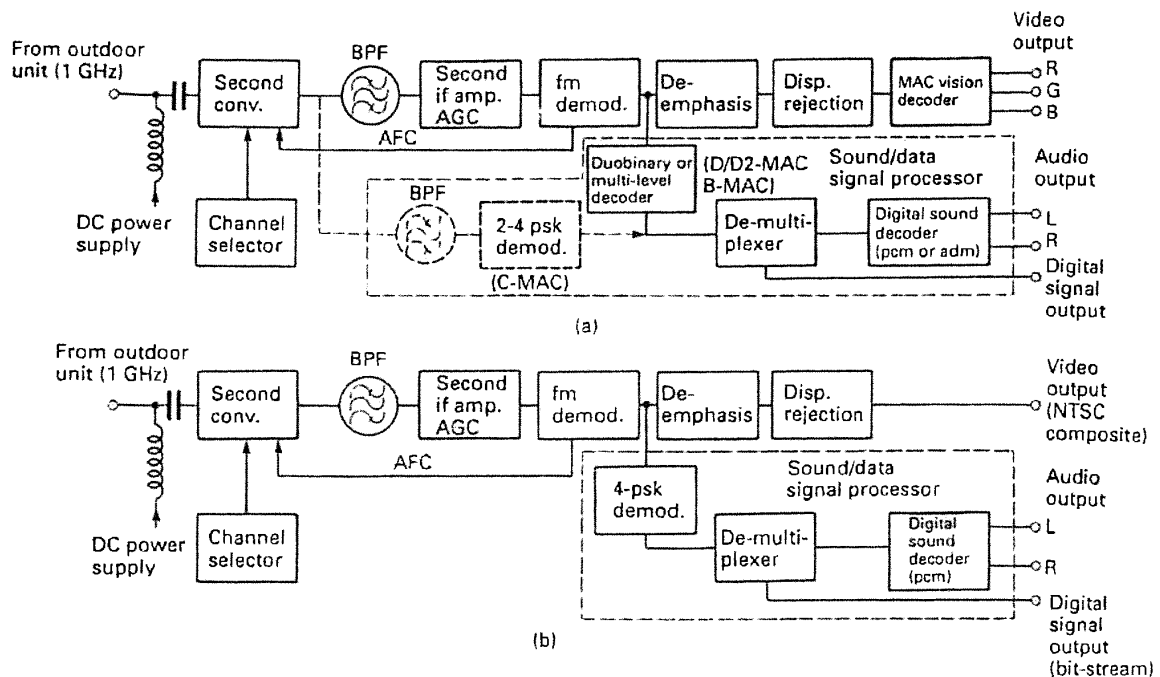


Figure 31.4 Indoor units: (a) MAC/packer family and B-MAC system, (b) digital subcarrier/NTSC system

The fact that the dbS receiver has been developed using these high technologies will enable it to play an important role as the centre of new broadcast receiving systems.

the 12 GHz band are shown in Table 31.2.

31.4 Requirements

Satellite broadcasting is capable of sending the same high quality TV signals to the home viewer as to the studio. These signals need to be available to the majority of the viewers in the service area, and the system should be harmonized with the existing and future systems. For any satellite broadcasting system, the receiver is a crucial element affecting as it does the system performance and cost. Important criteria for any receiver system include:

- low cost
- high performance
- unit interchangeability
- compatibility between dbS receivers and existing TV sets
- less spurious radiation
- protection against interference
- expandability to future satellite broadcasting systems
- ease of handling as a domestic appliance

Thus it is necessary to specify target performances for the receiving equipment and promote its standardization.

31.5 Operation and performance

31.5.1 Receiving antenna

Compared with the vhf signal, the shf (12 GHz) signal is extremely short in wavelength (about 2.5 cm) and may be captured by a parabolic dish.

The receiving antenna needs to have a gain that is sufficient relative to the required G/T (which indicates the sensitivity of the receiver — see section 31.5.1.2) in the receiving area, and it should be available at a moderate cost, be easily installed, and retain a stable performance through its life.

Three types of receiving antennas are shown in Figure 31.5. The receiving antenna with a parabolic reflector (dish) provides excellent gain and directivity.

Figure 31.6 shows the construction of a parabolic receiving antenna. The outdoor unit is directly connected to the primary feed to minimize loss.

The signals reflected by the parabolic reflector focus at the top of a primary feed. The signals collected by the primary feed are introduced into the outdoor unit through a polarizer which converts the polarization of the signals from circular to linear.

Antenna	
Type and Gain	Type 60 75 90 100 120 (dB) 34.5 36.5 38.0 39.0 40.0
Output VSWR	< 1.3
Output port	WRJ-120 waveguide/BRJ-120
Outdoor unit	
Input port	WRJ-120 waveguide/BRJ-120 flange with waterproof
Noise figure	< 4 dB
Input signal level	- 80 dBm \pm 10 dB/channel
Input vswr	< 2.5
Overall gain	48 \pm dB
1st local oscillator frequency	10.678 GHz \pm 1.5 MHz (-30° - 50°C)
leakage power	< -30 dBm
1st if	1.036 - 1.332 GHz
Indoor unit	
Input signal level	43 dBm +15 dB -18 dB
Input vswr	< 2.5
2nd if	either 134.26 or 402.78 MHz
Leakage power of 2nd local oscillator	< -55 dBm
AFC gain	within \pm 500 KHz against \pm 2 MHz deviation
Received signal quality	
frequency response	video: +1 dB for 50 Hz 4.2 MHz, within -3 dB at 4.6 MHz sound: within +1 - -3 dB at 15 KHz for A mode within +1 - -3 dB at 20 KHz for B mode
linearity	video: Differential Gain < 5% Differential Phase < 5°
bit error rate	< 3×10^{-4} at c/n = 9 dB, before error correction

(CCIR, Study Group, DOC, 10-11S/J-12, Japan)

Table 31.2 Characteristics of home receiving equipment

In the case of a flat antenna, the signals are received by several hundred small antenna elements on a board. Their currents are then collected and fed directly to the outdoor unit.

31.5.1.1 Antenna performance

Performance criteria for a dba antenna are gain, efficiency, G/T, directivity and vswr.

31.5.1.1.1 Gain and efficiency

Antenna gain, G_r , is given by:

$$G_r = \eta \left(\frac{4\pi A}{\lambda^2} \right) \quad (31.0)$$

where η is efficiency, A is aperture area and λ is wavelength. The efficiency η is the product of six specific efficiencies:

$$\eta = \eta_m \cdot \eta_g \cdot \eta_r \cdot \eta_b \cdot \eta_i \cdot \eta_s$$

where

- η_m = efficiency due to reflector material
- η_g = polarization efficiency
- η_r = efficiency due to accuracy of reflector surface
- η_b = efficiency due to blockage of radiated beam
- η_i = illumination distortion
- η_s = efficiency due to spillover at the surrounding area of reflector

The larger the aperture area is, the higher the gain. If the aperture area is constant, each efficiency must take a higher value in order to increase the gain of the antenna.

However, $\eta_i \cdot \eta_s$ has an optimum point, and its maximum value is about 0.8.⁵ An effective surface accuracy better than 0.5 mm rms is desirable under all weather conditions. When the accuracy of the reflector surface is 0.5 mm rms, η_r is 0.94, so that $\eta_r \cdot \eta_i \cdot \eta_s$ is 0.75. Therefore, the upper limit of the antenna efficiency is considered to be approximately 75 per cent.

Using an offset parabolic reflector, antenna efficiency of more than 70 per cent has already been achieved in commercial antennas.

31.5.1.1.2 Overall performance of receiver and G/T

The input carrier signal power, C_i , of the outdoor unit is expressed by:

$$C_i = \left(\frac{W G_t}{L_t L_p} \right) \left(\frac{\lambda}{4\pi d} \right)^2 \frac{G_r}{L_r} \frac{1}{L} \text{ watts} \quad (31.1)$$

where

- W = the transmitting power of the broadcasting satellite
- G_t = the transmitting antenna gain
- L_t = the feeder losses between the transmitter and transmitting antenna
- L_p = the pointing losses of the transmitting antenna
- λ = the wavelength of the carrier signal
- d = the distance between the satellite and the receiver
- G_r = the receiving antenna gain
- L_r = the feeder losses between the receiving antenna and the outdoor unit
- L = the rain and atmospheric losses

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The first term of the equation is the *equivalent isotropically radiated power* (eirp) expressed as follows:

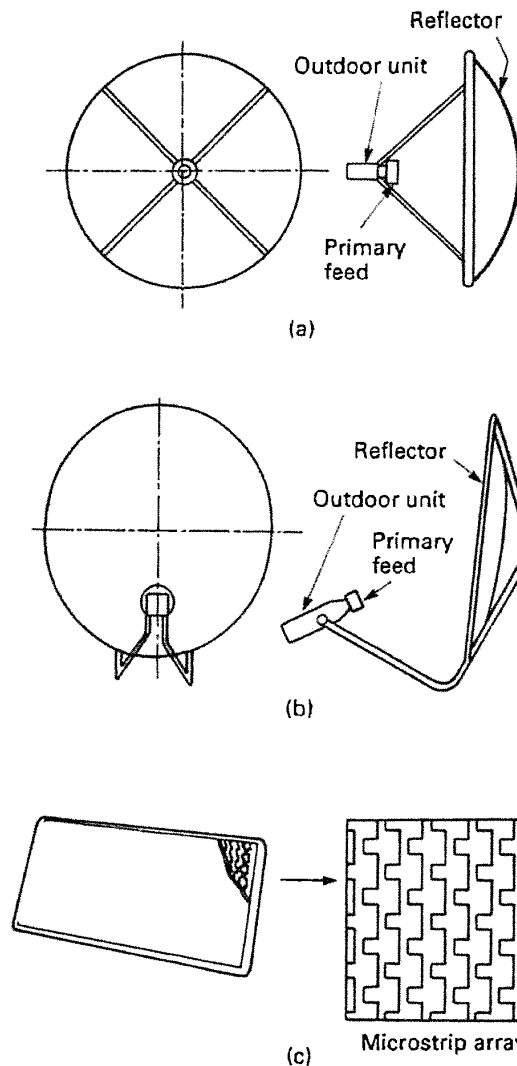


Figure 31.5 Types of receiving antenna: (a) centre feed parabolic antenna, (b) offset parabolic antenna, (c) flat antenna

$$\text{eirp} = \frac{W G_r}{L_r L_p} \text{ watts} \quad (31.2)$$

The second term is the *free space path loss* (L_r):

$$L_r = \left(\frac{\lambda}{4\pi d} \right)^2 \quad (31.3)$$

Therefore, when a receiving antenna with gain G_r is used, the *input carrier signal power* of the outdoor unit is expressed by:

$$C_i = \text{eirp} \cdot \frac{G_r}{L_r L_p} \text{ watts} \quad (31.4)$$

On the other hand, the input noise power of the outdoor unit includes the signal source noise (noise of antenna) and the noise generated in the receiver (converted value at the input port).

The *input noise power* of the receiver (outdoor unit) N_i is:

$$N_i = K T_a B + K T_0 B(n-1) = K B [T_a + T_0(n-1)] \text{ watts} \quad (31.5)$$

where K = Boltzmann's constant, $1.38 \times 10^{-23} \text{ W/Hz K}$

T_a = the noise temperature of the receiving antenna (K)

B = the bandwidth (Hz)

T_0 = the reference temperature (K)

n = the overall noise figure of the receiver, expressed as a power ratio

The expression $T_a + T_0(n-1)$ is the *system noise temperature*, T_s .

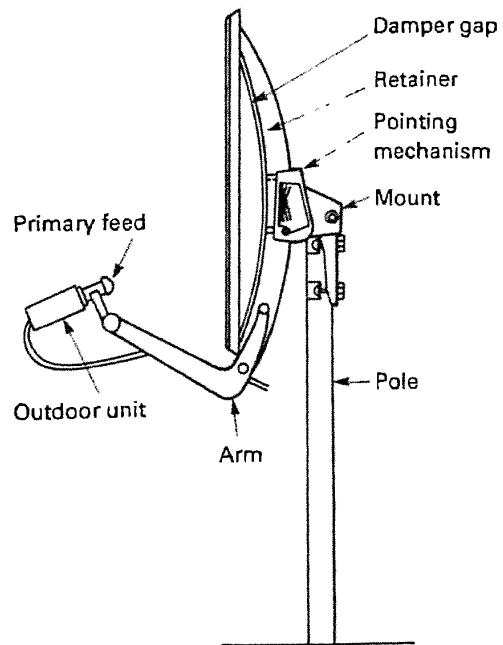


Figure 31.6 Construction of parabolic receiving antenna

The resultant carrier/noise ratio, C_i/N_i , at the input port of the receiver (outdoor unit) is calculated as follows:

$$C_i/N_i = 10 \log \frac{\text{eirp}}{K B [T_a + T_0(n-1)] L_r L_p} \text{ dB} \quad (31.6)$$

To attain excellent picture quality, a carrier/noise ratio of more than 14 dB is needed.

The *figure of merit* (G/T) is the receiving antenna gain (G_r) divided by the system noise temperature (T_s) a compound of the antenna and outdoor unit noise temperatures; it indicates the sensitivity of the receiver. From equations (31.4) and (31.5) the figure of merit is:

$$\frac{G_r}{T_s} = \frac{C_i}{N_i} \cdot \frac{L_r L_p K B}{\text{eirp}} \quad (31.7)$$

The figure of merit is defined in terms of the pointing error and the coupling loss by WARC-BS³ as follows:

$$G/T = \frac{\alpha \beta G_r}{\alpha T_a + (1-\alpha) T_0 + (n-1) T_0} \quad (31.8)$$

where α = the total coupling losses, expressed as a power ratio,
 β = the total losses due to the pointing error, polarization effect and ageing, expressed as a power ratio,
 G_r = the effective gain of the receiving antenna, expressed as a power ratio and taking account of the method of feeding and the efficiency,
 T_a = the effective temperature of the antenna,

T_o = the reference temperature (=290 K),
 n = the overall noise figure of the receiver, expressed as a power ratio.

In WARC-BS planning, the value of the G/T is 6 dB/K for individual reception, and 14 dB/K for community reception.

However, an antenna for home use needs to be small and lightweight. An antenna with a relatively small diameter is now capable of providing the required figure of merit by virtue of its low noise figure and high efficiency. Even an antenna of 60–70 cm diameter provides a G/T between 10 and 12 dB/K.

At 12 GHz, the most popular shape of antenna for dbS reception is one with a conventional or offset parabolic reflector. The choice of diameter depends on the required figure of merit.

The relationship between the antenna diameter and the effective figure of merit is indicated in Figure 31.7 (including the loss in gain introduced by the antenna pointing error).

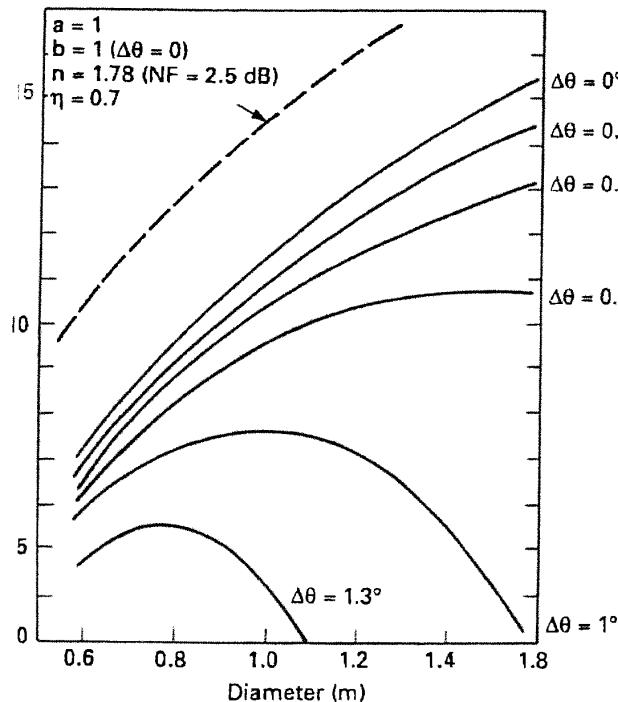


Figure 31.7 Relationship between antenna diameter and effective figure of merit. $a = 0.9$ b (considering losses due to overall pointing error $\Delta\Omega$). $T_a = 150$ K, $T_o = 290$ K, $n = 2.51$ ($nf = 4$ dB), $f = 12$ GHz, $\eta = 0.6$

31.5.1.1.3 Directivity

The directivity of the receiving antenna is expressed by the co-polar (same component as antenna polarization) and cross-polar (reverse component to antenna polarization) reference pattern. In order to avoid mutual interference between countries, WARC-BS has decided on the antenna reference pattern shown in Figure 31.8.

31.5.1.2 Reflector

The receiving antenna must provide not only high performance but also high reliability (including high resistance to wind), and ease of mass production and installation for domestic use.

Compared with a centre feed parabolic antenna, an offset parabolic antenna has the following advantages:

- There is no need to allow for a decrease in gain and degradation of the antenna pattern due to the scattering effect of the feed support and the blockage by the primary feed.
- It is possible to reduce the insertion loss by connecting the outdoor unit directly to the primary feed.
- Wind pressure is lower as it strikes a less curved surface. Adhesion of snow on the offset antenna is much less.
- It is easy to attach a cover to the outdoor unit to prevent temperature increases.

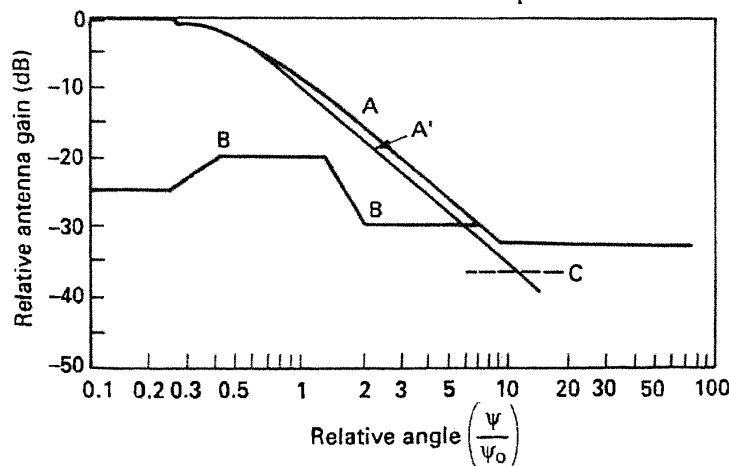


Figure 31.8 Co-polar and cross-polar reference patterns for receiving antennas (Regions 1, 3). Curve A: co-polar component for individual reception without side-lobe suppression; Curve A': co-polar component for community reception without side-lobe suppression; Curve B: cross-polar component for both types of reception; Curve C: minus the on-axis gain (WARC-BS)

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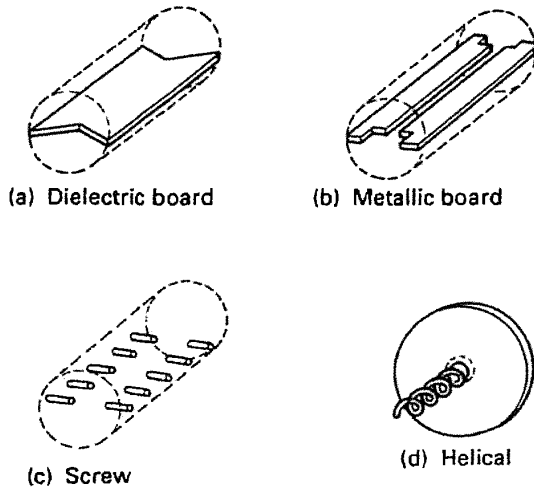


Figure 31.9 Circular polarizers

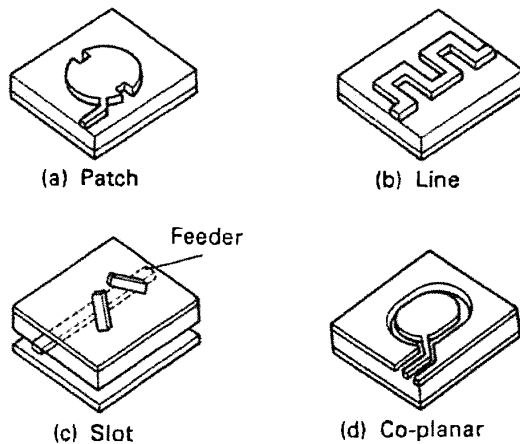


Figure 31.10 Patterns of microstrip antenna element

- It is easy to mass produce by pressing or plastics moulding techniques.

The reflector of the antenna is made of metal or reinforced plastics with conductive material such as embedded wire mesh, carbon fibre or evaporated aluminium.

The offset parabolic antenna has been proved to be able to offer high efficiency and reliability and be suitable for mass production.

31.5.1.3 Primary feed and polarizer

The primary feed is located at the focus of the reflector. The right-hand circular polarized signal from the satellite is reflected and changed to left-hand circular polarization by the reflector, then introduced into the primary feed. The primary feed with a circular polarizer converts the signal to linear polarization and feeds it to the input stages of the outdoor unit.

A circular polarized signal is considered to have two linear polarized components with a phase difference of 90° . The circular polarizer changes both linear polarized components to the same phase. The feed is usually horn type.

The various types of polarizers available are shown in Figure 31.9.

31.5.1.4 Flat antenna

A flat antenna receives signals from the satellite, using a flat board containing small built-in antenna elements instead of the reflector. Flat antennas are roughly divided into two types: *microstrip* and *slotted waveguide*. Recently, microstrip antennas using a printed circuit board have been developed and put on sale in Japan. In these, several hundred small antenna elements shaped on a board (antenna array) are combined and their currents are collected at the feed point. The microstrip element comes in various patterns as shown in Figure 31.10. The printed circuit board is made of a material such as teflon glass fibre.

The flat antenna using a simple microstrip has bigger feeder losses and narrower bandwidth, and so it is necessary to employ only the best material with a contrived antenna element pattern and a contrived construction of the board such as a multilayer construction in order to attain wideband and low-loss performances.

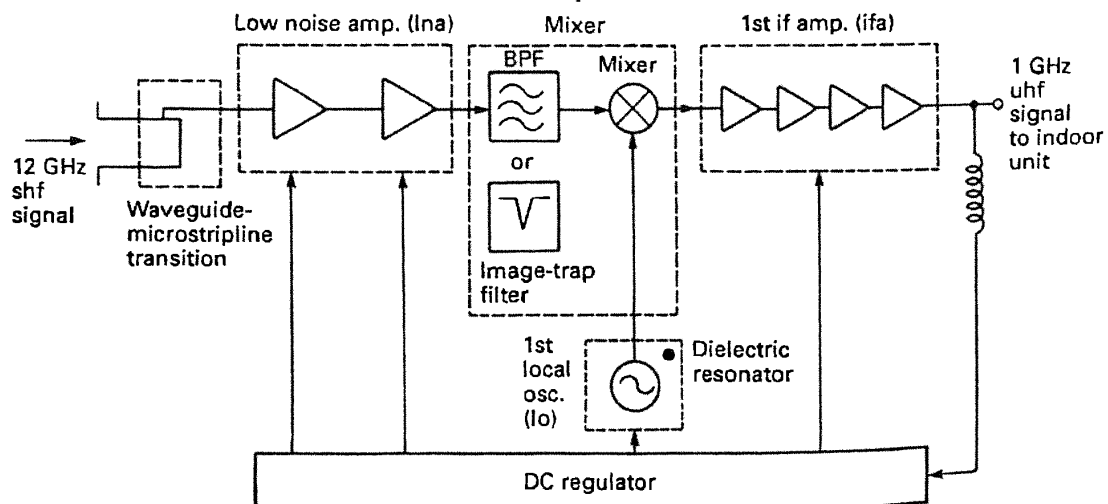


Figure 31.11 Outdoor unit (low noise converter)

The flat antenna is expected to be cheaper, easier to install and less obtrusive than the parabolic antenna. In addition, the flat antenna can incorporate an outdoor unit and has the capability of being pointed with electronic steering. However, the present flat antenna has the disadvantages of lower efficiency due to feeder losses and higher effective temperature of the antenna, particularly in a large unit. Further developments are necessary.

31.5.2 Low noise converter (outdoor unit)

The *low noise converter* (lnc) is one of the important components that determines the receiving signal quality of the dbS receiver. A block diagram is shown in Figure 31.11. The 12 GHz fm TV signals received are converted to the first if signals and then amplified in the uhf band. The lnc performs block-down conversion of the satellite TV signals to the first intermediate frequency range.

The signal collected by the antenna is introduced through the waveguide to the microstrip/line transition circuit. This circuit feeds the signals to the next low noise amplifier in microstrip/line mode. The lna is made of two to three stage gallium arsenide fets (mesfets). The output signal of the lna is mixed with the output of first local oscillator and converted to the first if signal.

A technology known as MIC (microwave integrated circuit) with low noise gallium arsenide fets and Schottky barrier diodes is mainly used to integrate the lnc circuits.

The output signal of the mixer is amplified by the first if stage. The dc power is supplied to the lnc by the indoor unit through the coaxial interconnecting cable.

To ensure reliability under severe environmental conditions, a hermetically sealed and waterproof construction is applied to the outdoor unit.

31.5.2.1 Waveguide microstrip/line transition

The signal collected by the feed device of the antenna is passed through the waveguide to the microstrip/line transition circuit, which is formed on the substrate.

Using the phenomenon of frequency cut-off in the waveguide, a high-pass filter is formed, which suppresses the image band. The first local oscillator frequency is usually allocated below the signal band. The image band is approximately 9–10 GHz, which may be used in a strong radar band. The outdoor unit with waveguide filter is effective for image suppression. Using this filter, image suppression of more than 50 dB is attained.

The waveguide microstrip/line transition circuit operates so

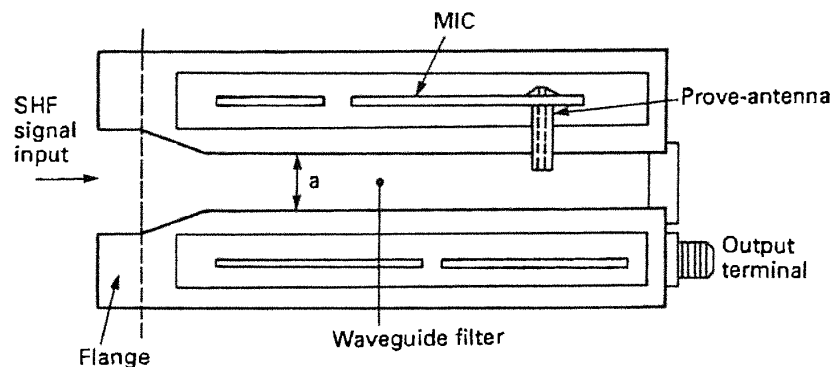


Figure 31.12 Principle of an outdoor unit

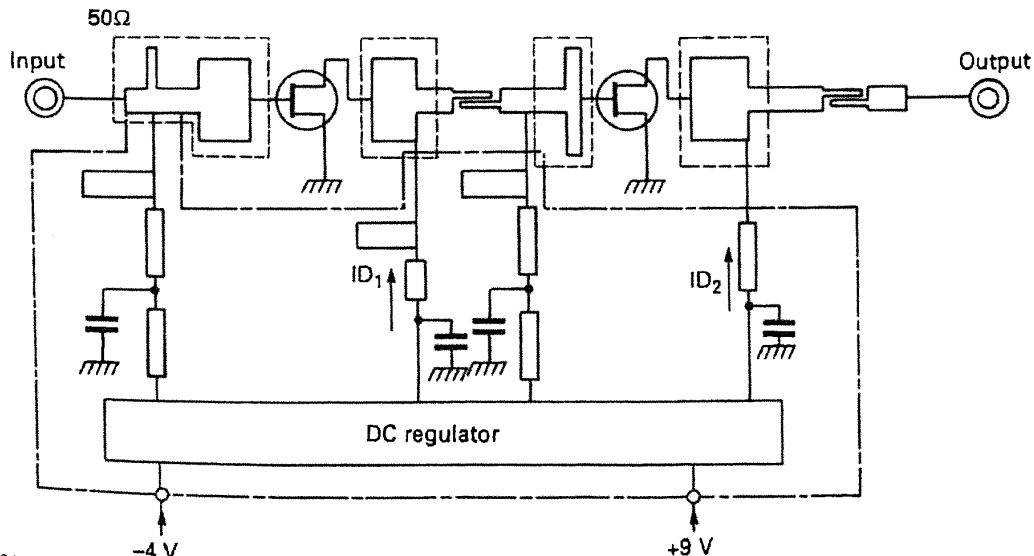


Figure 31.13 A low noise amplifier circuit. The broken line boxes are impedance matching circuits, designed to minimize the noise in the input and to maximize the gain in the output

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as to transmit the signals in the waveguide to the microstrip/line of the low noise amplifier. This circuit may be formed by inserting a probe-antenna connected to the microstrip/line into the waveguide.

31.5.2.2 Low noise amplifier

The low noise amplifier is composed of a two to three stage amplifier using low noise devices such as gallium arsenide fets (see section 31.5.2.2.2) or high electron mobility transistors (hemts — see section 31.5.2.2.3) and is fabricated by MIC technologies. Sufficient gain must be provided by the lna to overcome losses in the transmission from it to the following receiver circuit and to override noise which originates after the lna.

31.5.2.2.1 Noise and gain of low noise converter

The key parameter for a low noise converter is the noise performance. In order to receive and amplify the weak signal from a satellite minimizing the noise generated in the receiver, a low noise figure (nf) and high gain are required.

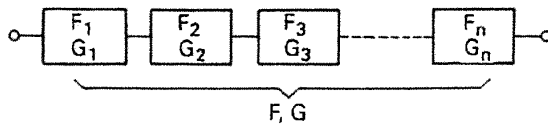


Figure 31.14 Low noise amplifier. F = noise figure, G = gain of amplifier

In the low noise amplifier shown in Figure 31.14, the overall nf is expressed by:

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots + \frac{F_n - 1}{G_1 G_2 \dots G_{n-1}} \quad (31.9)$$

Therefore, the overall nf is affected mainly by the noise figure of the first stage, and less by the noise figures of the succeeding stages depending on the gain of the preceding stage. Consequently, a high gain amplifier with low nf is required in the preceding stage.

In a practical lnc, two lossy components (a filter and a mixer) succeed the lna. Assuming the first stage nf to be low, and the filter and mixer nfs to be less than 0.1 dB, the lna gain should be more than 15 dB.

The gain associated with a gallium arsenide fet in practice is approximately 9 dB, and a two to three stage lna is necessary. The fet has 1.6 dB nf, and a typical noise factor for the lnc of 2.0 dB has been obtained. Table 31.3 shows an example of gain and nf allotment of a low noise converter.

	NF (dB)	Power gain
WG WM	0.1	—
LNA	1.7	18.0
MIX	5.0	— 5.0
IFA	2.5	35.0
Overall	2.01*	48.0

* Overall NF

$$NF = 0.1 + 10 \log \left(10^{1.7/10} + \frac{10^{(5+2.5)/10} - 1}{10^{18/10}} \right) = 2.01(\text{db})$$

Table 31.3 Typical gain and nf allotment of a low noise converter

The gain diagram is indispensable in designing a high performance lnc. The lnc gain is determined after the consideration of the noise factor of the indoor unit (which will not affect the receiver's nf) and the linearity of the lnc. Maximum interconnecting cable loss is about 15 dB for a 30 m cable. Assuming the noise factor of the indoor unit to be 12 dB, that of the lnc to be 3 dB and the gain of the lnc to be 48 dB, then the total increase will be 0.02 dB. This is a negligible figure.

LNC gain linearity is usually measured by third order intermodulation. The co-channel protection ratio is required to be 31 dB by WARC-BS. Assuming the intermodulation product of the input signals to be 24 dB, the third order intermodulation measured at the output port should be less than -55 dB. An overall gain of 48 dB is adequate for the lnc.

31.5.2.2.2 Gallium arsenide fet

The GaAs mesfet (gallium arsenide metal semiconductor field effect transistor) is an important device for the 12 GHz band lnc. The electron mobility of gallium arsenide is about six times that of silicon. The construction of a GaAs mesfet is simple but it has high-frequency performance superior to a silicon transistor in:

- decrease of the parasitic capacitances by isolation effect using a layer of semi-insulating GaAs
- decrease of the parasitic serial resistances and increase of the transconductance, g_m by a high electron mobility using an active layer of n-GaAs

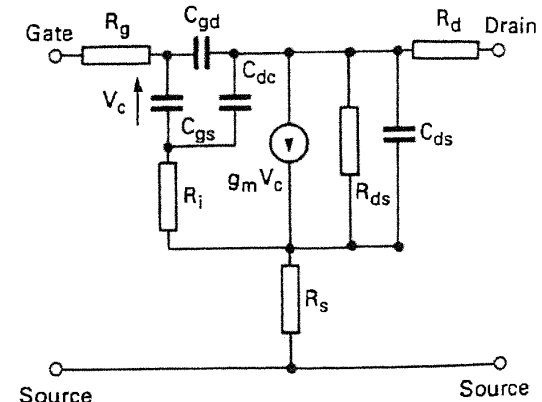


Figure 31.15 Equivalent circuit of GaAs fet

$$F_0 = 1 + 2 \pi K_f C_{gs} \sqrt{\frac{R_g + R_i}{g_m}} \quad (31.10)$$

where F_0 is the optimal value of noise figure, K_f is the fitting factor and f_o the operating frequency

To achieve low noise performance, the epitaxial layer under the gate must be etched into a recess, and the gate length must be short. This requires the use of an ion-implantation technique.

Characteristics of low noise GaAs fets in the shf band have been greatly improved in recent years through the progress of semiconductor technologies.

A GaAs fet with a very low noise figure has become available at reasonable cost. Using low noise GaAs fets with a noise factor of 1.6 dB, a typical overall nf of about 1.9 dB has already been achieved in the receivers on sale.

31.5.2.2.3 HEMT

In addition to the improvements of GaAs mesfets, an epoch-making new device, hemt (high electron mobility transistor), has recently been introduced into the line of the dbS receiver.

In 1978, it was discovered by Dingle and others of Bell Labs that the hetero-structure (junction of different semiconductors) of AlGaAs-GaAs has a high electron mobility. A hetero-structure has about double the electron mobility of a Schottky-barrier junction as is formed in the GaAs mesfet. The GaAs fet with a hetero-structure is a *hemt*.

The hemt is fabricated by a similar process to the GaAs fet. Because of its superior transconductance, the hemt has an extremely low noise performance.

With a hemt having a noise factor of 1.2 dB, a typical overall nf of about 1.5 dB has been achieved in receivers.

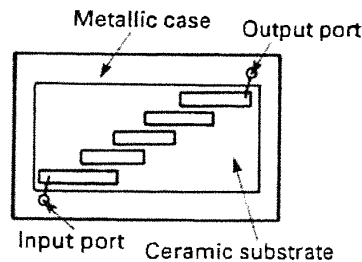


Figure 31.16 Construction of a band-pass filter

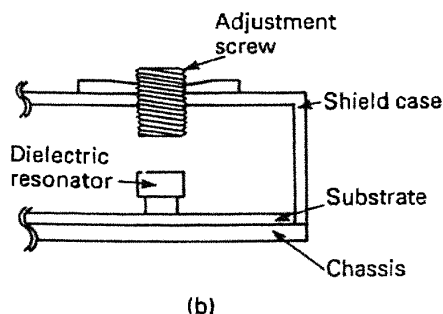
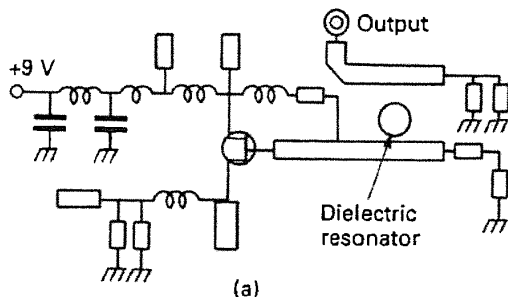


Figure 31.17 First local oscillator: (a) circuit, (b) adjustment mechanism

31.5.2.2.4 MIC

The receiver front end consisting of a low noise amplifier, local oscillator and mixer, is fabricated by the hybrid integrated

circuit technology (called *microwave integrated circuit*) whereby the fets and the associated passive elements formed by microstrip/lines and stubs are mounted or formed on the same teflon or ceramic substrate. Figures 31.13 and 31.17 show examples of the MIC. A more integrated version will eventually be possible, and microwave monolithic integrated circuits (e.g. active and passive components integrated with the same GaAs chip) may be foreseen.

31.5.2.2.5 Band-pass filter

A band-pass filter has two objectives: image suppression, and spurious radiation control. A typical filter is shown in Figure 31.16. The image suppression could be kept below 40 dB only by this filter. Instead of using the band-pass filter, an image-trap filter is used in the image recovery mixer (see section 31.5.2.4) for image suppression.

31.5.2.3 First local oscillator

The first local oscillator is a simple GaAs fet oscillator stabilized by a dielectric resonator. The oscillator principle is usually either reflection or feedback oscillation. A GaAs fet used in the oscillator is sometimes a medium power unit. Figure 31.17 shows an example of a first local oscillator circuit using a common source fet and adjustment mechanism.

Frequency drift of the local oscillator should be within the frequency space. A practical figure will be within 1 MHz, with the provision of afc (automatic frequency control) in the indoor unit. Using a dielectric resonator, stable oscillator frequency can be obtained in whatever severe thermal condition the outdoor unit may be placed.

31.5.2.4 Mixer

The 12 GHz TV signal is mixed with the first local oscillator signal, and converted to the first if signal (1 GHz). The mixer usually utilizes a balanced diode mixer, which effectively reduces local oscillator leakage. An alternative mixer is single ended with image recovery, which reduces insertion loss. The image-trap filter suppresses image signals and reflects image components generated by diode mixing. Image components are put into the diode mixer again and improve the conversion efficiency. Figure 31.18 shows the image recovery mixer. Mixer diodes are GaAs Schottky barrier diodes. Silicon Schottky barrier diodes are also useful for the mixer.

31.5.2.5 First if amplifier

The first if amplifier amplifies 1 GHz uhf band signals, and drives the signals into the connecting cable. A significantly high output power stage is necessary in order to reduce the output stage distortions and/or intermodulations. The first if amplifier with or without gain adjustment and tilt-compensator consists of a four to five stage amplifier using a low noise high frequency transistor, and its gain is commonly 30–40 dB.

31.5.3 Indoor unit

The indoor unit has channel selection and fm demodulation functions (see section 31.3.4). The video signals and the sound/data signals are derived by demodulation fm signals. In the C-MAC system, however, the sound/data signals are directly demodulated by a 2-4psk demodulator from rf signals. The derived sound/data signals are processed by demodulator and/or decoders to the baseband audio signals. The MAC vision signals are also decoded by a MAC vision decoder to baseband signals.

As the dbS signal format is different in each country, so the indoor unit has to be designed specifically for each.

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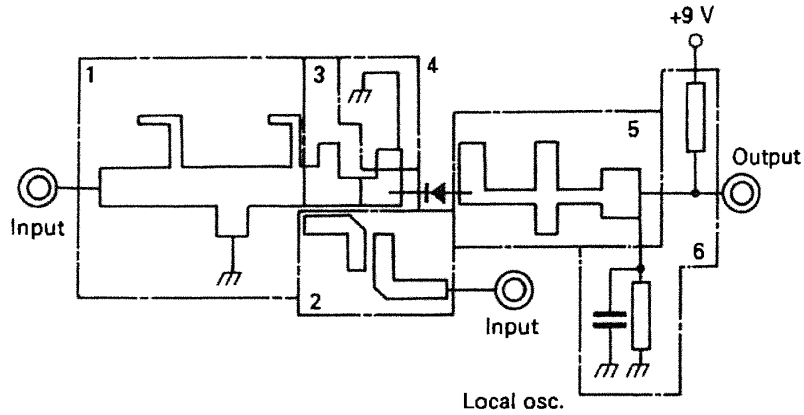


Figure 31.18 Mixer circuit: 1 image-trap filter, 2 band-pass filter, 3 impedance matching circuit, 4 short circuit for if signal, 5 low-pass filter, 6 bias circuit

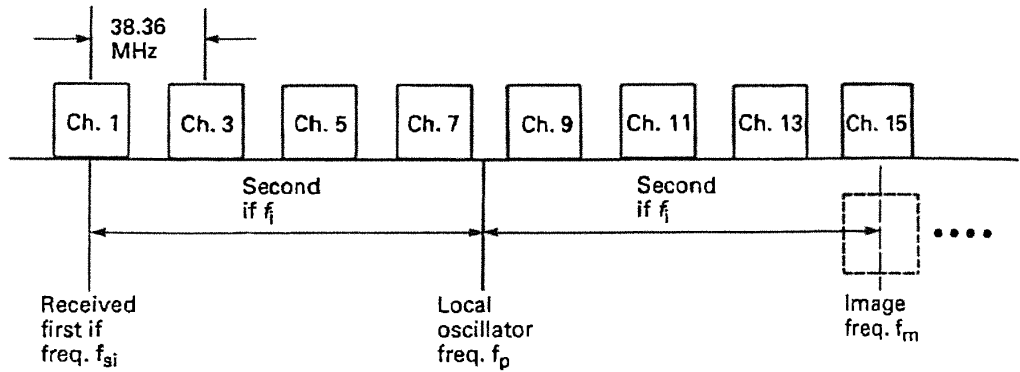


Figure 31.19 Relationship between second if and other frequencies

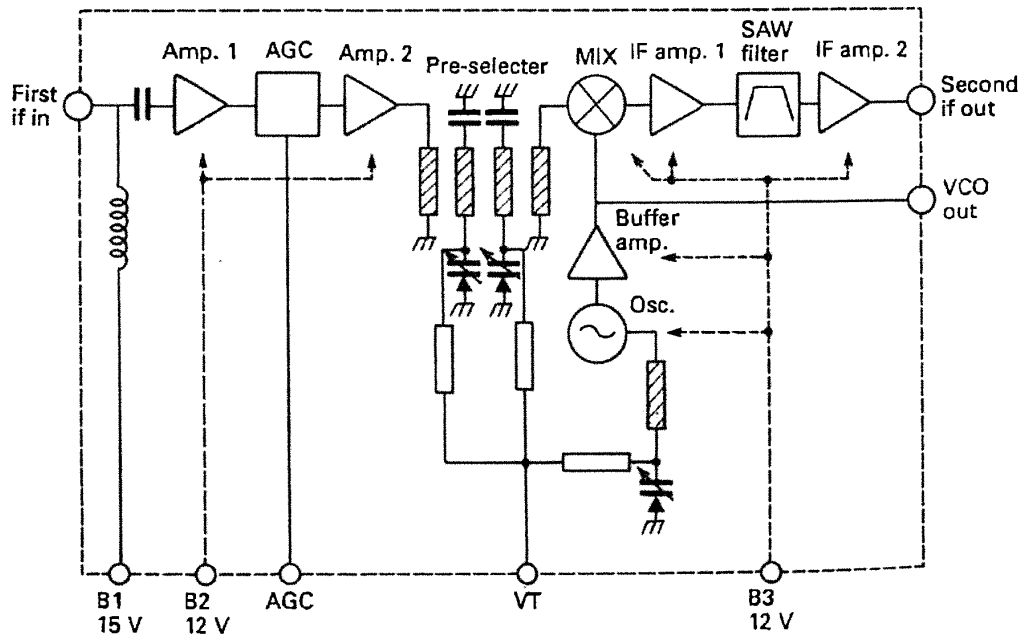


Figure 31.20 Configuration of second converter

31.5.3.1 Second intermediate frequency and bandwidth

The 12 GHz dbS receiver employs a double frequency conversion system.

The second is freely selectable, but one must take into account high power terrestrial broadcasting transmitters, and interference from the second local oscillator radiation of other receivers. For community reception, a common outdoor unit is connected to multiple indoor units. When the selected value of the second is smaller than the value of the total frequency bandwidth allocated to satellite broadcasting, local oscillator frequencies fall in part of the first if band. To keep this interference as low as possible by arranging the second local oscillator frequencies between any two adjacent channels allocated to that area, the following relationship is desirable for the second if:

$$f_i = F(n + 1/2) \text{ MHz}$$

where F is 38.36 MHz for Regions 1 and 3 or 29.16 MHz for Region 2.

The relationship is valid when the selected frequency f_i is smaller than the value of the total frequency bandwidth in that area. However, if $2f_i$ is less than the total bandwidth, there is the possibility of image frequency interference, and insertion of a tunable band-pass filter may be advisable before the mixer.

The second intermediate frequency bandwidth is defined by WARC-BS to be 27 MHz (Regions 1 and 3) or 24 MHz (Region 2). Filter characteristics outside the band should be specified as co-channel protection ratio defined by WARC-BS.

31.5.3.2 Second converter

The second converter is one of the most important parts in an indoor unit, and is similar to the TV tuner, especially the uhf tuner.

It converts the uhf intermediate signal to a second if signal with a 27 MHz or 24 MHz bandwidth, set by its channel selection switch. The second converter consists of a first if amplifier with agc, a tunable preselector, a mixer, a local oscillator with buffer amplifier and a second if amplifier which includes a surface acoustic wave (saw) filter.

In order to control the gain of the second converter and minimize the intermodulation, an agc using PIN diodes is provided in the first if amplifier.

In its rf stage, a tunable preselector (band-pass filter) is preferable because of the reduction in image rejection and the reduction in the second local oscillator radiation and IM3. The tunable preselector, which is composed of a double tuned band-pass filter varicap diodes and controlled by the voltage tracking the second local voltage controlled oscillator, reduces image interferences effectively.

31.5.3.3 Second if amplifier

In the second if amplifier, filtering, gain control and limiter functions are provided.

The band-pass filter is used to suppress the out-of-band noise. The overall selectivity of the receiver is set by this filter with a bandwidth of 27 MHz or 24 MHz. The filter consists of a saw filter or a helical filter. The saw filter assures a high quality picture, with no need for adjustments, suppressing the out-of-band noise and giving superior phase delay characteristics. The helical filter is used in high second intermediate frequencies.

Gain control is provided for stabilizing the if signal level of the demodulator to obtain stable demodulation characteristics, and keep intermodulation distortion at low level.

The limiter suppresses undesirable am noise over a wide carrier frequency and improves the c/n threshold characteristics. In a phase locked loop (pll) demodulator, the limiter function is included. (Fig. 31.22)

31.5.3.4 FM demodulator

The fm demodulator converts the incoming frequency modulated signal to a baseband signal containing a video signal and digitally modulated sound/data signals. It can generate an afc signal.

The fm demodulator of the dbS receiver requires wide-band demodulation. There are two types of demodulators, the double-tuned discriminator and the phase locked loop demodulator shown in Figure 31.21. A pll type demodulator is most suitable for a dbS receiver because of its high performance. It has the advantage of s/n improvement in the picture

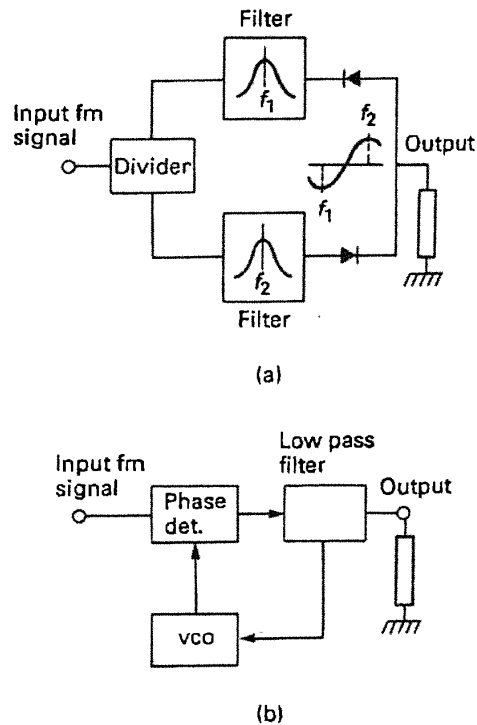


Figure 31.21 FM demodulator: (a) double-tuned discriminator, (b) phase locked loop demodulator

under low c/n condition, known as *threshold extension*, simple adjustment and excellent linearity. Basically, it consists of a phase detector (or mixer), low-pass filter (or loop filter) and voltage controlled oscillator (vco). When the pll is locked to the incoming fm signal, the demodulated signal can be derived from the output of the loop filter. The performance of the pll is determined by the time-constant of the loop filter.

A pll fm demodulator improves the threshold c/n by about 2-3 dB.

31.5.3.5 Energy dispersal signal rejection

The energy dispersal signal has a triangular waveform (a half, or a quarter, of the field frequency); it is added to the video signal in order to avoid concentration of the energy spectrum in fm transmission. The rejection circuit removes this triangular waveform; unless it is removed, there will be flicker interference on the picture. A high-speed peak clamping circuit or a pulse clamping circuit (Figure 31.22) is used.

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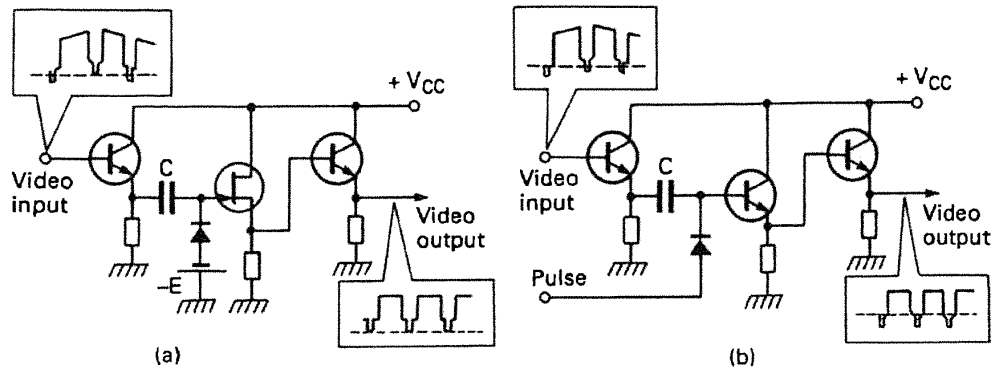


Figure 31.22 Examples of analogue energy dispersal rejection circuits: (a) peak clamp, (b) pulse clamp

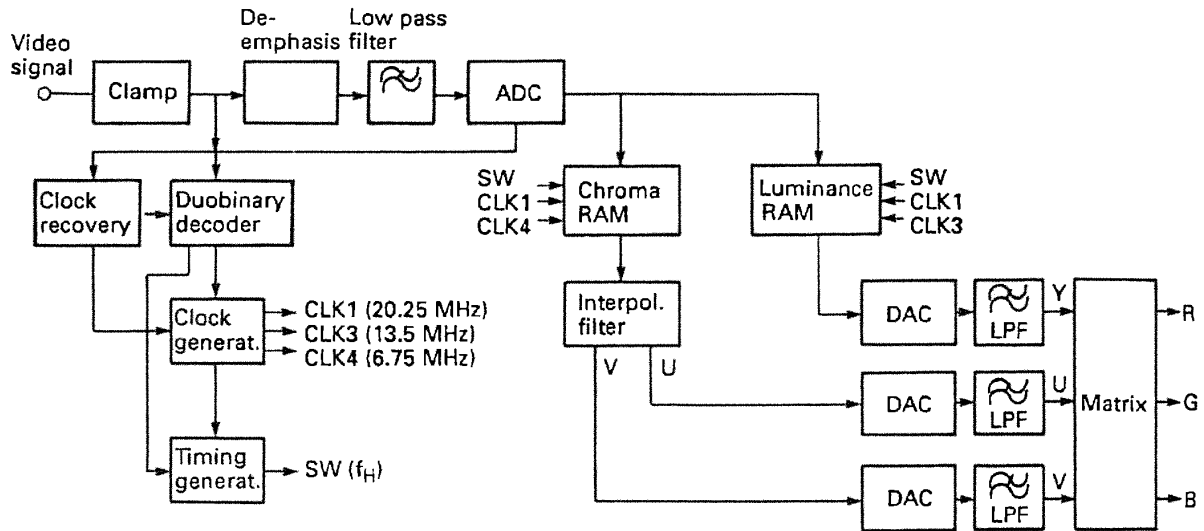


Figure 31.23 Notional diagram of a typical MAC vision decoder

31.5.3.6 De-emphasis

As the signal/noise ratio of the high-band video signal is degraded in fm transmission, the video signal is pre-emphasized in high-band, before transmission. So, the received video signal is de-emphasized by means of a network having a reverse characteristic to the pre-emphasis. The characteristic of the de-emphasis is defined by each transmission system.

31.5.3.7 Video signal processor

The digital subcarrier/NTSC system employs the conventional NTSC composite video format, so this system does not particularly require video signal processing in the dbS receiver. As the MAC/packet family employs time division multiplexing, it is necessary to decode the MAC vision signals into conventional video signals in the receiver. The MAC vision signal processor is described in the following section.

31.5.3.7.1 MAC vision decoder

The present vision standard specifications of the MAC/packet family is compatible in each system, so the circuit configuration

of the MAC vision decoder is the same. Figure 31.23 is a notional diagram of a MAC vision decoder. The output signals of the fm demodulator are applied to the MAC decoder through the clamp, de-emphasis, and low-pass filter.

On the other hand, clock signals and timing pulses are generated from the data burst signals, which are multiplexed during the line blanking interval.

The MAC vision signals are digitized by an ADC using sampling clock frequencies (20.25 MHz for C/D and D2-MAC and 21.477 MHz for B-MAC, 525-line systems). They are divided into two signals: one for the luminance and one for the colour-difference signal processing circuit (comprising line memories). The time expansion of the luminance signal, Y, is achieved by writing the digitized vision signals into a line memory at the above sampling frequency and reading them out at a frequency of 13.5 MHz (for C/D and D2-MAC) or 14.318 MHz (for B-MAC 525-line). This can be done by two line memories used alternately for read and write.

The colour-difference signals are sequentially transmitted within alternate lines as U and V. U is transmitted on odd active lines and V on even active lines. A chrominance signal is transmitted one line before the associated luminance signal.

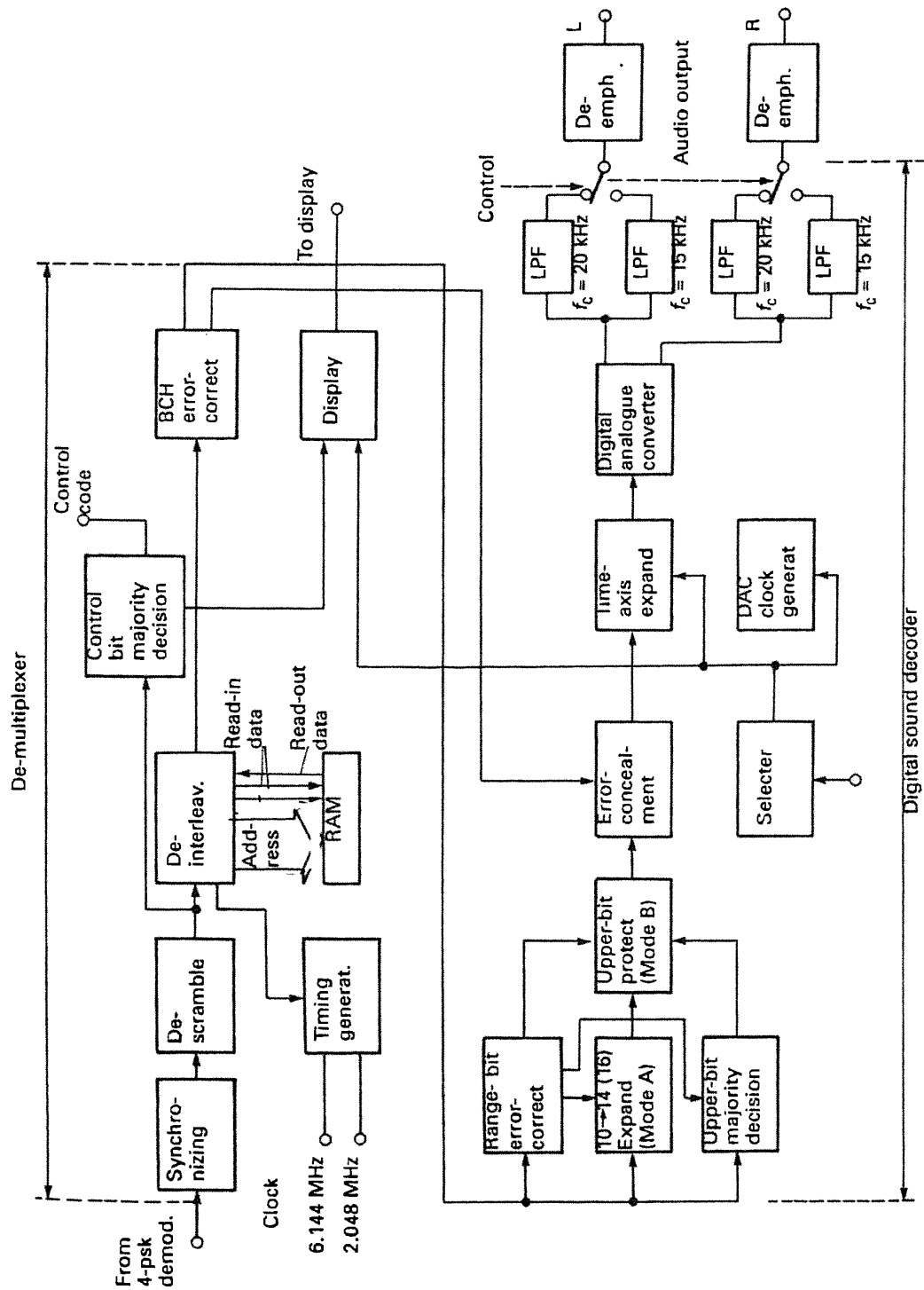


Figure 31.24 Sound/data signal processor (digital subcarrier/NTSC). LPF = low pass filter

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Symbol		Phase-change	Rest state
y_n	x_n		
0	0	0°	0
0	1	$+90^\circ$	1
1	1	$\pm 180^\circ$	2
1	0	-90°	3

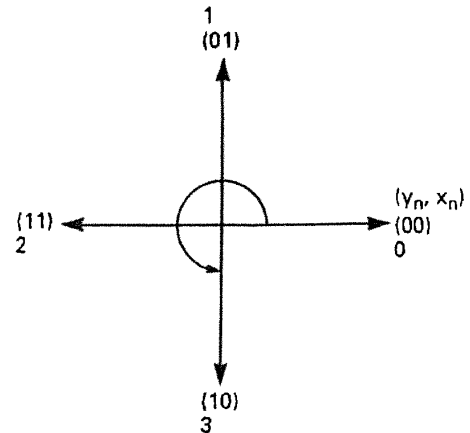


Figure 31.25 Transition rule of quadrature phase shift keying (4psk)

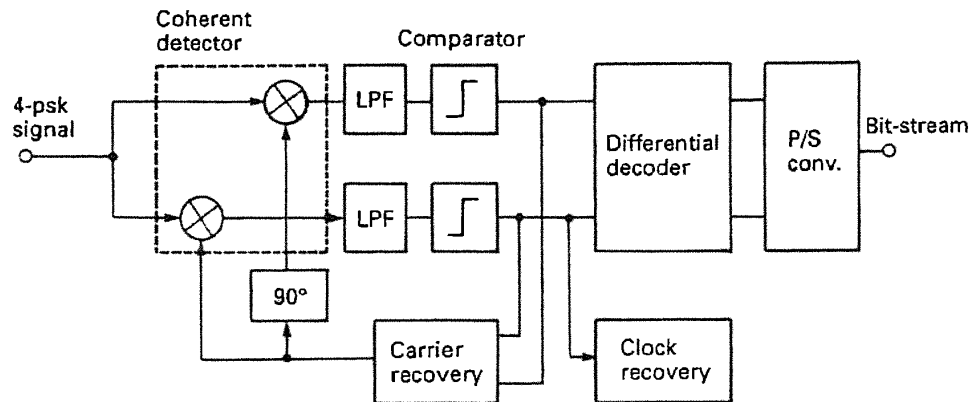


Figure 31.26 4psk demodulator (synchronous detection)

Due to sequential transmission, the colour-difference signal processing necessitates digital interpolating processing of two adjacent chroma lines. For interpolating processing, a 1:2:1 interpolation filter using a mean value of two adjacent chroma lines is employed.

The colour-difference signal processing is achieved in a similar manner to the luminance signal processing.

Y signals derived from the luminance signal processing circuit and U and V signals derived from the chrominance processing circuit are converted to analogue signals by a DAC, and are filtered to suppress the out-of-band spurious signals. Analogue signals of Y, U and V are converted by a matrix circuit to R, G and B signals.

31.5.3.8 Sound/data signal processor

The functions of a sound/data signal processor in the receiver comprise:

- demodulation or decoding of multiplexed sound/data signals
- transmission de-multiplexing of sound data and additional data
- digital sound decoding

For the dbx systems, sound signals are transmitted in coded forms together with the data signals and in the format of either continuous and rigid structure or flexible and packet structure.

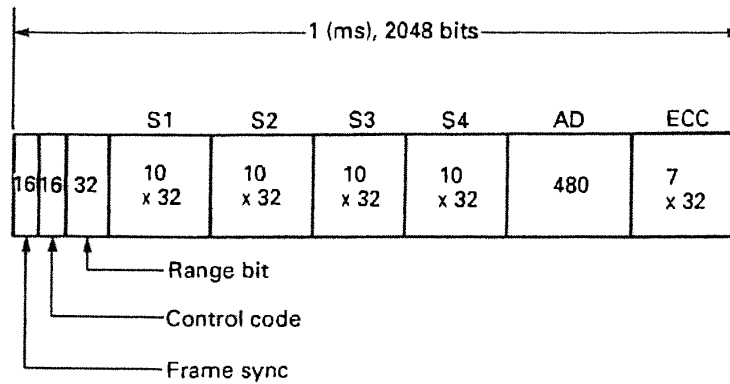
In *de-multiplexing*, the data for particular services are processed and selected from the demodulated or decoded sound/data signals.

The modulation, multiplexing and coding forms of the sound/data signal are different in each transmission system. The sound/data signal processor (comprising demodulator/decoder, de-multiplexer and digital sound decoder) utilized in the receivers of each system are described in sections 31.5.3.8.1–31.5.3.8.3.

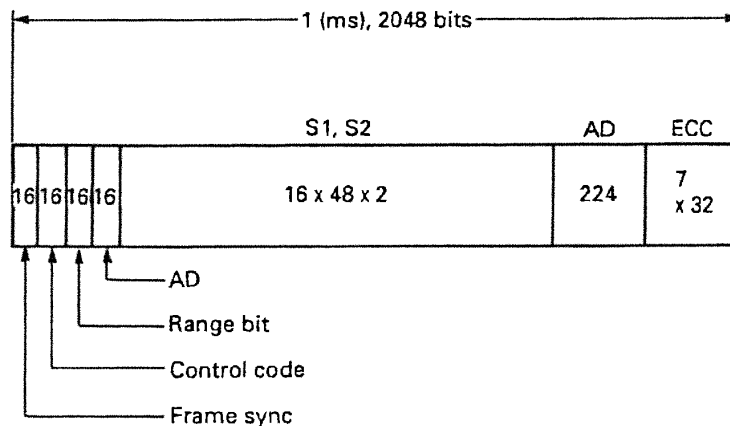
31.5.3.8.1 Digital subcarrier/NTSC system

The sound/data signals are carried with a 5.73 MHz subcarrier (4psk modulated), which is frequency-multiplexed with the conventional NTSC vision signal. In the receiver, the subcarrier signal is separated by a 1 MHz band-pass and fed to a 4psk demodulator. The sound/data capacity (mean data rate) of the system is 2.048 Mbit/s. A continuous and rigid structure of sound/data multiplexing is employed. In the case of data transmission, however, the multiplexing employs the packet format in an additional data area.

In this system, the digital sound coding uses pulse code modulation. The system can provide four high quality audio channels (*mode A*) of 15 kHz bandwidth with near instantaneous 14/10 bit companding, or two very high quality audio channels (*mode B*) of 20 kHz bandwidth with linear 16 bit coding. Figure 31.24 shows a sound/data signal processor of the digital subcarrier/NTSC system.



(a)



(b)

Figure 31.27 Sound/data frame structure (digital subcarrier/NTSC system): (a) mode A, (b) mode B; S_n = audio signal n , AD = additional data, ECC = error correction code

The 4psk (*quadrature phase shift keying*) is a four-state modulation method converting logic changes in signal pairs of bits (called *symbols*) of the bit-stream to the phase-change (rest states 90° apart) of the carrier signal by the rule shown in Figure 31.25. Because the phase/changes between one symbol and the next (differential logic) are transmitted, the receiver can easily recover them without detecting the absolute carrier phase.

For the 4psk demodulation, synchronous detection is mainly utilized as shown in Figure 31.26. In this method, the coherent detector, composed of two multipliers, performs a phase detection of an incoming 4psk signal by using two reference carriers, whose phases differ by 90° . Since the carrier of the incoming signal is suppressed, carrier recovery is first required. For synchronous detection, the carrier recovery is conducted by a phase locked loop circuit. To conduct the carrier recovery

using pll, it is necessary for the output of the coherent detector to eliminate the rest-state phase shift components of the incoming carrier. For this purpose, methods such as Costas loop, remodulation and others are employed.

After the phase lock detection, the demodulated signals are discriminated and converted into two bit-streams by the comparator. A clock recovery circuit using a pll is also provided and regenerates a bit-synchronizing clock signal from the bitstream. The output of the 4psk demodulator is applied to a differential decoder having reverse logic to transmitted differential logic symbols. The output of the differential decoder, which has parallel two bit-streams, is applied to the parallel/serial converter and converted into a serial bit-stream.

The *de-multiplexer* in the receiver converts a bit-stream into a sequence of sound samples and data. It comprises such circuits as a frame sync circuit, a descrambler, a de-interleaver, a

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control code decoder and error correction circuit. (See Figure 31.24.) Figure 31.27 shows a frame structure.

When the bit-stream is applied to the de-multiplexer, first a frame synchronization word of 16 bits is detected by a frame sync circuit from the bit-stream and is synchronized. The frame sync circuit has a flywheel effect so as not to mis-synchronize in the case of mis-detection or error occurrence.

In the descrambler, a pseudo-random binary sequence (prbs), which is previously added to randomize the transmission spectrum and to recover easily the bit-clock signals, is eliminated from the bit-stream by adding (modulo 2) the same prbs.

Descrambled sound/data signals are divided into control code decoder and de-interleaver. Error protection of control codes are performed by 8/15 majority decision of each 15 frames.

Interleaving is previously applied to the bit-stream in order to error-correct effectively, converting burst errors which occur in transmission to random errors. In this system, the 2048 bits of each frame are interleaved with a distance of 32 bits. At the RAM of the de-interleaver, bit read-in (transmission sequence) and read-out are performed with the sequence shown in Figure 31.28.

After the de-interleaving, error correction of BCH(63,56) is performed. The BCH(63,56) code makes one bit error correction and two bits error detection for one subframe of 63 bits.

BCH is from the inventors Bose, Chaudhuri and Hocquengham. (63,56) means that the total bit numbers of the block is 63 bits, and its information bit number is 56 bits. (The check code bit number is 7 bits.)

After the error correction, the sound-sample words are decoded by using sound coding law. Digital companding

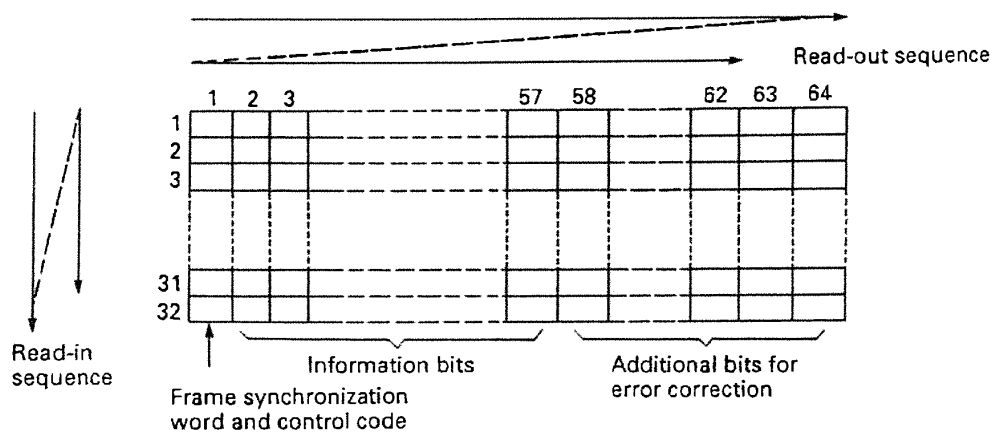


Figure 31.28 Bit interleaving sequence (digital subcarrier/NTSC system)

Range of largest magnitude "a"		Range number	Sound-sample bit number													
Positive	Negative		MSB	2	3	4	5	6	7	8	9	10	11	12	13	LSB
$a < 2^{13}$	$ a \leq 2^{13}$	0	* 0 1													
$a < 2^{12}$	$ a \leq 2^{12}$	1	* 0 1	0 1												
$a < 2^{11}$	$ a \leq 2^{11}$	2	* 0 1	0 1	0 1											
$a < 2^{10}$	$ a \leq 2^{10}$	3	* 0 1	0 1	0 1	0 1										
$a < 2^9$	$ a \leq 2^9$	4	* 0 1	0 1	0 1	0 1	0 1									

* Positive = 0
Negative = 1

Parts of the same bits as MSB bit

Bits used for companded code

Extracted bits

Figure 31.29 Near instantaneous companding

techniques are used to produce efficient coding formats for broadcast audio transmission.

In mode A of the digital subcarrier/NTSC system, *near instantaneous companding* is adopted in order to reduce the sound-sample word from 14 bits to 10 bits (as shown in Figure 31.29). Near instantaneous coding is done by using five coding ranges of compression to which each 32 successive samples in a period of 1 ms has been subjected. In the receiver, in order to reproduce original sound-sample words, range-codes are error corrected and sample words are expanded using range-codes and companding law.

After the error correction, an error concealment of a sound-sample word using error flags added in the error correction procedure is performed by the interpolation of two adjacent sound-samples. The error-concealed sound data are expanded along the time axis by DAC interfaces and are converted by a DAC to analogue audio signals.

31.5.3.8.2 MAC/packet family

The digital sound/data signals are multiplexed during the line blanking interval. In the receiver, the sound/data signals are derived and demodulated by a 2-4psk demodulator (in C-MAC), or decoded by a duobinary decoder (in D and D2-MAC). The data capacities (mean data rate) of each system are 3.08 Mbit/s in the C and D-MAC systems and 1.54 Mbit/s in the D2-MAC system, respectively. For the MAC/packet family, a data burst of each line is defined to contain 206 bits in the C and D-MAC systems and 105 bits in the D2-MAC system, respectively.

Using these bits, various forms of sound/data signal, synchronization word and control code are transmitted.

In the MAC/packet family, the digital sound coding adopts pcm. The systems are capable of providing eight high quality audio channels of 15 kHz bandwidth with near instantaneous 14/10 bit companding.

Figure 31.30 shows a sound/data signal processor of the MAC/packet family.

For the C-MAC/packet system, multiplexing is carried out at rf on the modulated signals. The sound/data signal is modulated using 2-4psk.

The 2-4psk is a modulation method converting logic 1 to a phase change of $+90^\circ$ and logic 0 to -90° . The phase-change of the carrier between adjacent bits is only 90° (Figure 31.31).

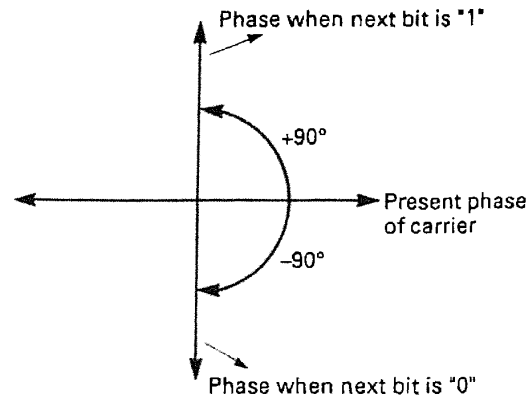


Figure 31.31 The rule of 2-4psk

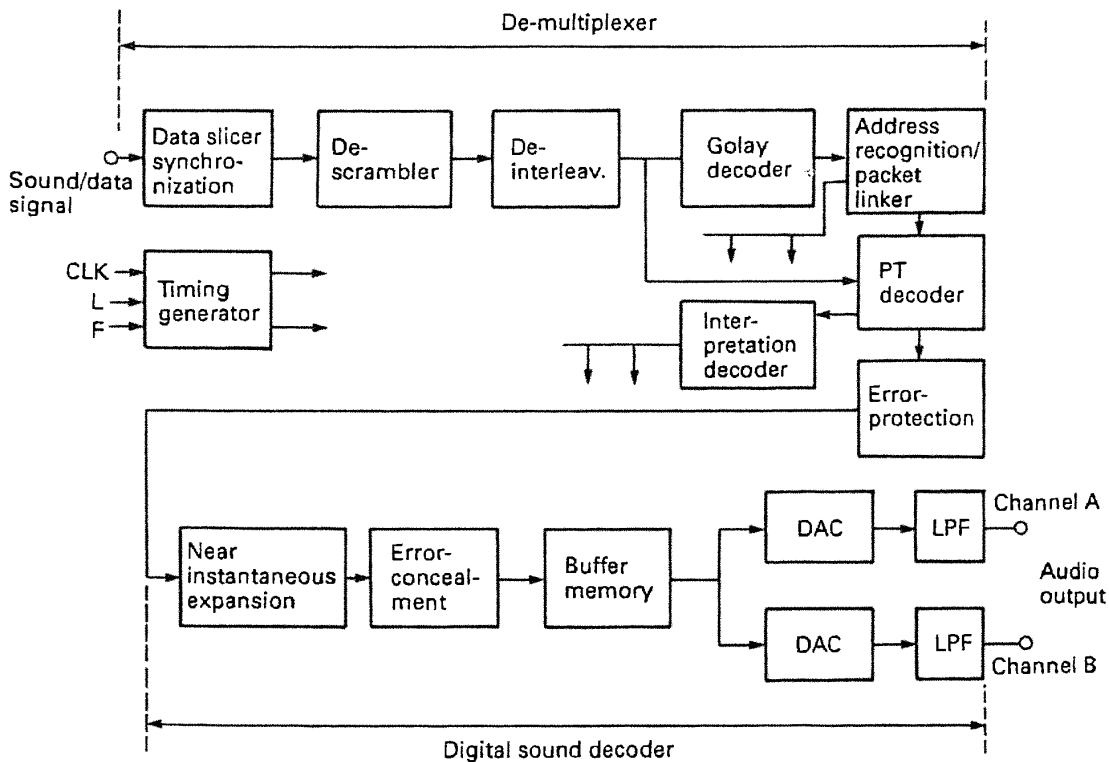


Figure 31.30 Sound/data signal processor (MAC/packet family)

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The demodulation of the 2-4psk signal is accomplished by using methods such as differential demodulation and coherent demodulation.

The principle of differential demodulation is shown in Figure 31.32. The incoming signal through a band-pass is split into two paths. One is phase-shifted and delayed by one bit, then the two signals are multiplied together. The sound/data bit-stream can be derived from the result.

For the D and D2-MAC/packet systems, sound/data multiplexing is carried out at the baseband during the line blanking interval. The sound/data signal is coded in duobinary form. With this coding method, the signal has three characteristic levels where the extreme levels (+1, -1) represent logic 1 and the intermediate level represents logic 0. An example of duobinary transcoding is illustrated in Figure 31.33. By using duobinary coding, the sound/data signals can be transmitted with reduced bandwidth compared with nrz.

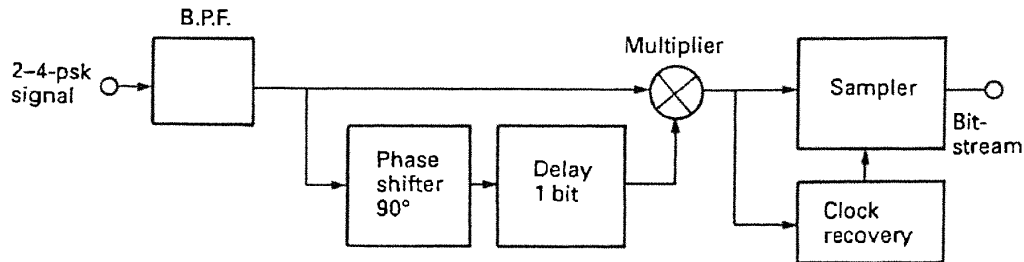


Figure 31.32 2-4psk differential demodulator

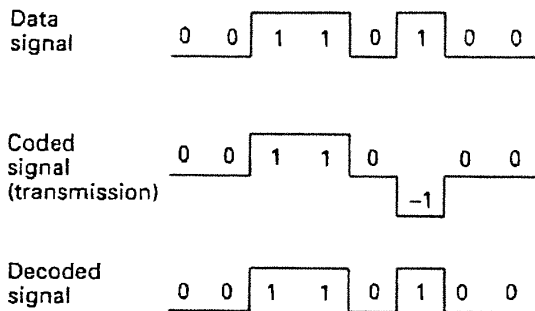


Figure 31.33 Example of duobinary transcoding

The decoder can be a combination of a comparator and a transcoding logic.

The *de-multiplexer* in the receiver converts a bit-stream and all types of sound packet into a sequence of sound-sample words and data.

The sound/data of the MAC/packet family are transmitted by packet multiplexing. All the packets have the same length and each of them is composed of two parts: header and useful data.

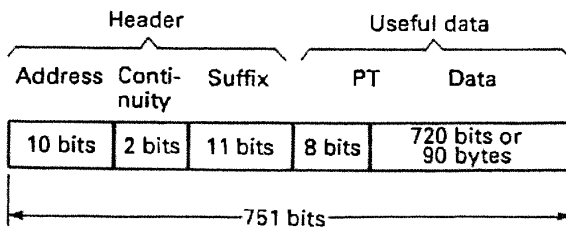


Figure 31.34 Packet structure

Selection of a particular service in the receiver is carried out by recognizing the address of the desired service carried in the packet header. A continuity index is used to assure the link between successive packets of the same service. The pt (*packet type*) code informs the sound decoder of the nature of the sound signal such as the coding law and audio bandwidth.

The de-multiplexer comprises a sync circuit, a de-scrambler, a de-interleaver, a Golay decoder, a pt decoder, an address recognition/packet linker and an error-protection decoder (see Figure 31.30).

The demodulated or decoded sound/data signals are applied to the de-multiplexer. Synchronization can be achieved either by *line sync word* (lsw) or by *frame sync word* (fsw). For sound/data synchronization, the start of the first packet is defined in line 1.

De-scrambling and de-interleaving are performed in the same way as in the digital subcarrier/NTSC system. In this

system, the 751 bits of each packet are interleaved with a distance of 94 bits.

The de-interleaved signal is divided into two paths, via a Golay decoder and a pt decoder. At the Golay decoder, the header is protected by using Golay cyclic code. At the pt decoder, the pt code is also detected and decoded. The output of the Golay decoder is applied to the address recognition/packet linker. The header and the pt code recognize and decide the required service of the received sound/data signal and control the functional configuration of the following process. After address recognition/packet linking, error-protection is performed. The error-protection method of the MAC/packet family operates at two levels: 1 parity bit/sample and Hamming 5 bit/sample.

For transmission, *near instantaneous companding* is done with compression from 14 bits to 10 bits per sample in 32 sample (1 ms) blocks. All the samples in each block are coded with coding ranges of compression determined by the magnitude of the largest sample in the block, and a scale-factor code is formed to convey the degree of compression to the receiver.⁹

In the receiver, sound-sample words are expanded from 10 bits to 14 bits using the scale factor and the near instantaneous companding law.

After the error correction, sound-samples with uncorrectable errors are concealed by interpolating two adjacent sound-samples.

31.5.3.8.3 B-MAC system

The sound/data signals are multiplexed during the line blanking interval. In the receiver, the sound/data signals are derived and decoded by a multi-level decoder. The data capacity (mean data rate) is about 1.6 Mbit/s. A data burst of each line is defined to contain 100 bits (50 symbols), and a packet structure is not employed.

The B-MAC system adopts an *adaptive delta modulation* (adm)⁸ as the sound coding method, and is capable of providing six high quality 15 kHz audio channels.

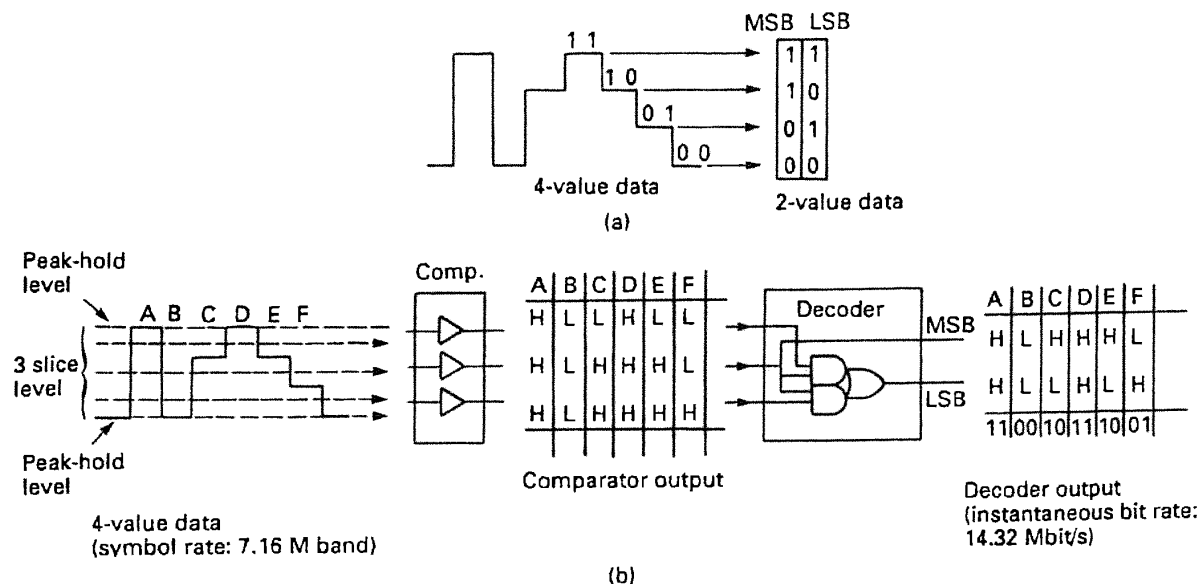


Figure 31.35 4-2 value transcoding: (a) transcoding rule, (b) operation

The sound/data signal is coded in *multi-level code*. Four level symbols are employed, each carrying 2 bits of information. Certain control data are transmitted using a two-level symbol which is a subset of the four-level symbols.

The decoder can use methods consisting of a peak-hold, comparators (or slicers) and a transcoding logic. Figure 31.35 demonstrates the principle.

ADM is a digital sound coding technique based on delta modulation and has a quality at least as high as an equivalent pcm system. This system offers high quality audio performance at the relatively modest bit rate per channel of 220-350 kb/s.

31.6 IC development

In order to achieve high performance, mass production and reasonable cost, it is important to introduce integrated circuits into the dbx receiver.

Monolithic IC development steps for the dbx receiver have been undertaken in the following areas:

- **microwave:** The monolithic microwave IC approach is considered to be one of the better methods of producing cost-effective and reliable microwave devices with accurate configuration in large quantity
- **second if:** the monolithic analogue ICs make possible high performances, minimize components and avoid adjustments
- **baseband signal:** analogue and digital lsi technologies can realize complex, large-scale and high performance signal processors in simple configuration

At present, various types of IC for the dbx receiver have already been developed and some of them have been used in the receivers on the market.^{7,9-13}

31.7 Re-transmission of satellite broadcasting

For maximum diffusion of satellite broadcasts, it is necessary to establish community reception systems or cable re-transmission

systems to allow numbers of households to receive satellite broadcasting services.

For the dbx re-transmission, the D and D2-MAC/packet systems were developed to have media transparency. Duobinary data transmission makes it possible to convert the fm satellite channels to the vestigial-sideband amplitude-modulated cable channels with reduced channel spacing (of at least 10.5 MHz for D-MAC, or 7 MHz for D2-MAC).

Generally, the cable re-transmission systems take one of two approaches:

- transmit fm signals with wide bandwidth, without changing the rf modulation parameters of satellite broadcasting (suitable for a small-scale system)
- transmit signals by remodulating to am compatible with the terrestrial broadcasting system (suitable for a large-scale system)

31.8 Future expansion

The broadcasting satellite will not only supply television satellite broadcasting but will also serve as the channel for important new broadcasting media.

The expected developments include high quality multi-channel digital sound broadcasting, conditional access broadcasting using scrambled signals, high definition TV, data broadcasting such as facsimile, teletext and still-picture broadcasting. To realize these broadcasting systems, additional reception equipment will be needed. It may be said that the satellite broadcasting systems have just moved from the phase of development of basic transmitting and receiving systems to the phase of development of new broadcasting systems using the satellite broadcasting media.

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