Chapter 4. AMPLITUDE MODULATION SYSTEMS

Classically, we have used analog communication systems to transmit information-bearing base-band signals from a point in space to another point in space or many points geographically distributed over a continuous medium of transmission, i.e., analog channels. They are primarily twisted-cable pairs, coaxial lines, and radio waves, all of which have different optimum pass-band characteristics and a carrier frequency w_c. Therefore, we need to develop various modulation schemes which operate efficiently in each of these channels.

Depending upon the application we upload the contents of the information-bearing base-band signal in the frequency-domain to the allocated frequency neighborhood using analog modulation systems. Some examples of base-band signals, their typical frequency ranges, typical application, and the band-pass ranges are tabulated in Table 4.1.

Table 4.1 Modulation Frequency Ranges and Applications

Information Signal	Frequency Band Name	Typical Application	Bandwidth	Base-band Frequency Range	Pass-band Frequency Range
Analog Speech/Audio	AM Band	AM Radio	10 kHz	0 - 4.0 kHz	530 kHz - 1.6 MHz
Analog Speech/Audio	FM Band	FM Radio	100 kHz	0 - 10.0 kHz	88 - 108 MHz
Analog Video/TV	VHF/UHF Bands	VHF/UHF Broadcast	6 MHz	88 - 108 MHz	30 - 300 MHz
Bipolar PCM Data	Digital Base-Band	T-1; T-5 Wired Telecom Network Infrastructure	1.544 MHz	0 - R_B Hz R_B : Bit Rate	1.544 - 260 MHz

4.1 BANDPASS SIGNALS and LOW PASS (BASEBAND) EQUIVALENTS

In all analog pass-band communications, a parameter of the sinusoidal carrier, $A.Cos(w_c t + \theta)$ is varied with a function of the modulating signal *x(t)*. The analog modulating system, its base-band and pass-band forms are tabulated in Table 4.2.

Table 4.2 Modulation Types and Functional Forms

Modulation Type	Baseband Functional Form	Baseband Functional Form
AM	$A(t) = f[x(t)]$	$A(t).Cos(2\pi f_c t + \theta_0)$
FM	$w(t) = f[x(t)]$	$A_c.Cos(2\pi f(t).t + \theta_0)$
PM	$\theta(t) = f[x(t)]$	$A_c.Cos(2\pi f_c t + \theta(t))$

In Table 4.2, A is the magnitude or magnitude function when it is time-dependent, $A(t)$ when it is time-dependent, the $f(.)$ stands for functional behavior. A_c is the amplitude of the carrier signal, $w(t)$ and $\theta(t)$ designate time-varying frequency and phase information, respectively.

4.1.1 Bandpass Signal: It is defined in the frequency-domain as:

$$V_{bp}(f) = 0 \quad for\,|\,f\,| < f_C - W \quad and\,|\,f\,| > f_C + W \tag{4.1a}$$

where $f_C = w_C / 2\pi$ is the carrier (modulation) frequency and W is the bandwidth of the low-pass equivalent of the signal $v(t)$. In addition to their time-domain waveforms, bandpass signals are normally described in terms of amplitude and phase plots as shown in Figure 4.1.

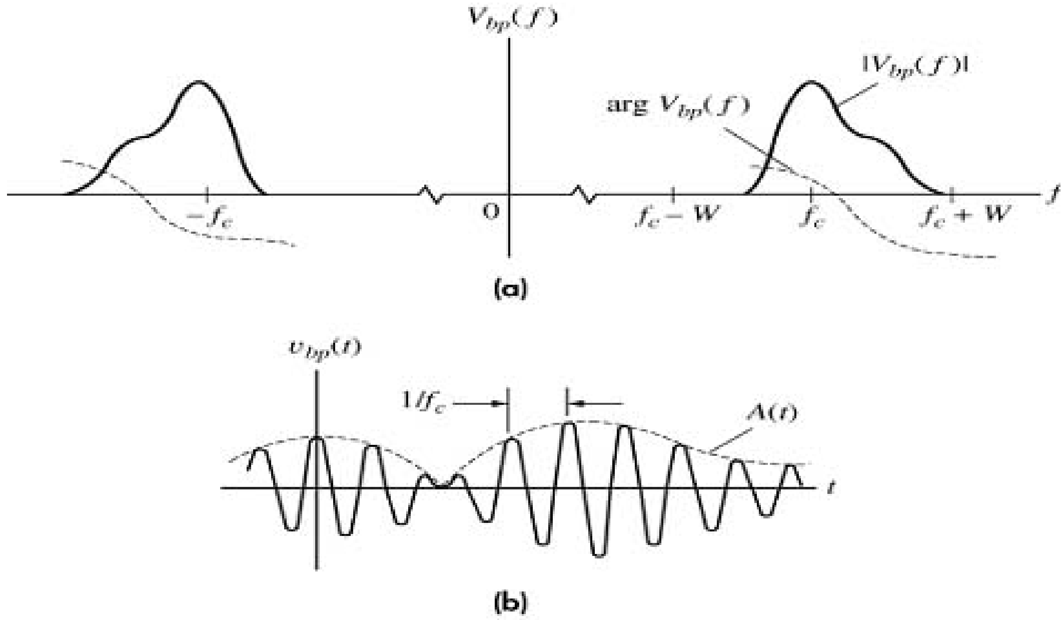


Figure 4.1 Bandpass signal illustrations in the frequency- and time domains. (Carlson p.144)

In analog communication tasks, we normally modulate signals with sinusoids and the corresponding form is given by:

$$v_{bp}(t) = A(t).Cos[w_C t + \phi(t)] \tag{4.1b}$$

where $A(t)$ is slowly varying information carrying signal, also called ***envelope*** and $\phi(t)$ represents the phase in the signal. It is common practice represent this equation in terms of complex-plane vectors as shown in Figure 4.2.

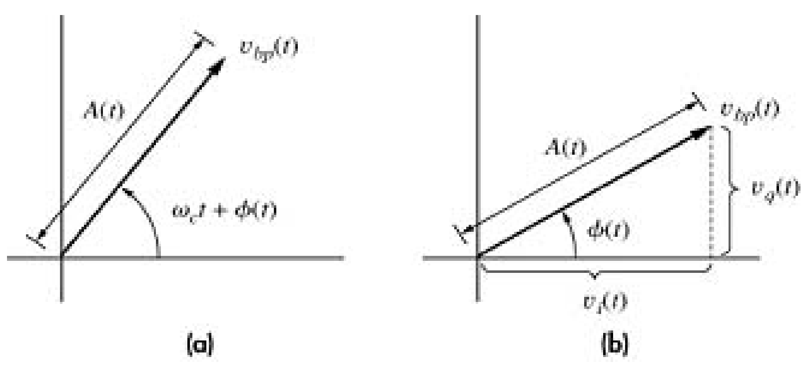


Figure 4.2 Phasor representation of a bandpass signal and "in-phase" and "quadrature" projections. (Carlson p.145)

In this case, bandpass signal is written in terms of "in-phase" and "quadrature" components as defined by:

$$v_i(t) \equiv A(t).Cos\phi(t) \quad and \quad v_q(t) \equiv A(t).Sin\phi(t) \qquad (4.1c)$$

$$v_{bp}(t) = v_i(t).Cosw_C t - v_q(t).Sinw_C t = v_i(t).Cosw_C t + v_q(t).Cos(w_C t + \pi/2) \qquad (4.1d)$$

It is not difficult to write in terms of amplitude and phase terms:

$$A(t) = \sqrt{v_i^2(t) + v_q^2(t)} \quad \text{and} \quad \phi(t) = \arctan\frac{v_q(t)}{v_i(t)} \qquad (4.2a)$$

In the frequency-domain, we have:

$$V_{bp}(f) = \frac{1}{2}[V_i(f - f_C) + V_i(f + f_C)] + \frac{j}{2}[V_q(f - f_C) + V_q(f + f_C)] \qquad (4.2b)$$

where $V_i(f)$ and $V_q(f)$ are the Fourier transforms of $v_i(t)$ and $v_q(t)$, respectively and they are low-pass signals with property:

$$V_i(f) = V_q(f) = 0 \quad for\ |f| > W \qquad (4.2c)$$

4.1.2 Low-pass Equivalent Signal: It is related to a bandpass signal shifted back to the origin (low-pass), which corresponds to the positive-frequency portion of the $V_{bp}(f)$ translated back down to the origin as shown in Figure 4.3.

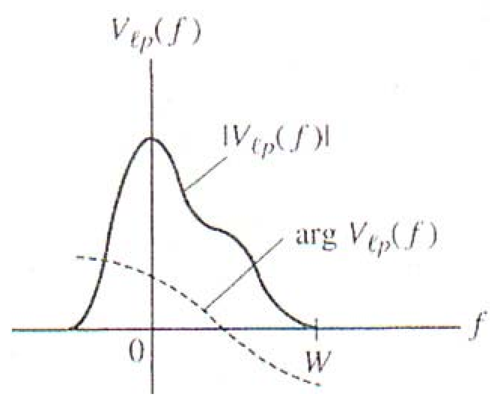


Figure 4.3 Low-pass equivalent of a bandpass signal (Carlson p.146)

$$v_{lp}(t) = F^{-1}\{V_{lp}(f)\} = F^{-1}\{V_{bp}(f - f_c).u(f + f_c)\}$$
$$= \frac{1}{2}.A(t).e^{j\phi(t)} \qquad (4.3a)$$

and the corresponding bandpass signal in terms of the low-pass equivalent version becomes:

$$v_{bp}(t) = 2.\text{Re}\{v_{lp}(t).e^{jw_C t} \text{ and } V_{bp}(f) = V_{lp}(f - f_C) + V_{lp}^*(-f - f_C) \qquad (4.3b)$$

4.1.3 Transmission in the Bandpass: The response of the transmitting channel can be studied either in the bandpass domain as shown Figure 4.4 (left) or in an easier fashion in the baseband (low-pass equivalent) domain (right).

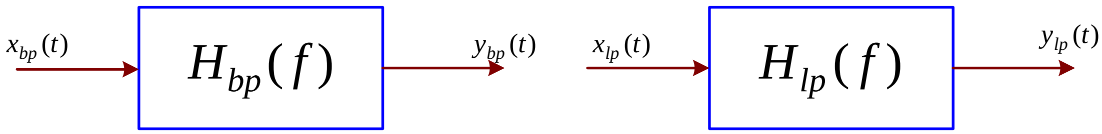


Figure 4.4 .System response in the passband (bandpass) and low-pass equivalent.

$$Y_{lp}(f) = H_{lp}(f).X_{lp}(f) \tag{4.4a}$$

where the low-pass equivalent transfer function is simply:

$$H_{lp}(f) = H_{bp}(f + f_C).u(f + f_C) \tag{4.4b}$$

the output in the time-domain is obtained simply from the inverse Fourier transform:

$$y_{lp}(t) = F^{-1}\{H_{lp}(f).X_{lp}(f)\} \tag{4.4c}$$

4.2 Suppressed Carrier Amplitude Modulation (DSB, AM-SC)

Amplitude modulation (AM) is the process of changing the amplitude of a relatively high frequency carrier signal $v_c(t)$ in accordance with the amplitude of the modulating signal (information) $x(t)$. It is a relatively inexpensive, low-quality form of modulation that is used for commercial broadcasting of audio and video signals. A generic block diagram for the modulator and the demodulator processes are shown in Figure 4.5.

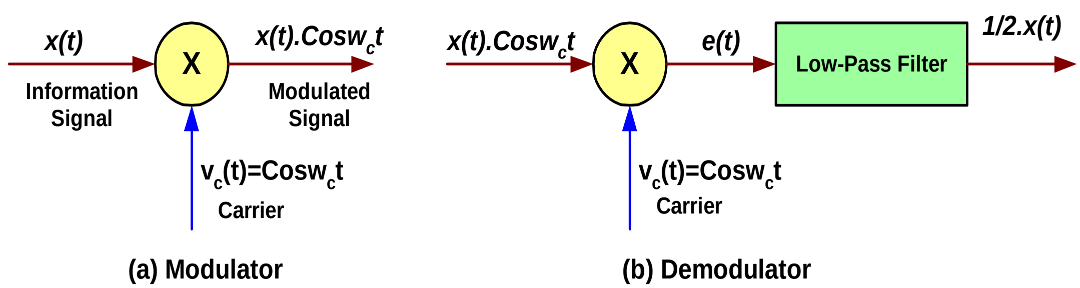


Figure 4.5 AM Modulator and Demodulator functional diagrams.

4.2.1 Modulation Process: From the modulation property of Fourier analysis we have the input-output relationships of the AM modulator in the time-domain and the frequency-domain:

$$s(t) = x(t).v_c(t) \tag{4.5a}$$

$$S(w) = \frac{1}{2\pi} X(w) * V_c(w) \tag{4.5b}$$

In the case when the carrier is a pure sinusoidal signal with an amplitude voltage level V_c and a frequency f_C Hz. We can write it as:

$$v(t) = V_c.Cos(2\pi f_c t) \tag{4.6}$$

The output corresponding to this sinusoidal carrier from the modulator is simply:

$$s(t) = V_c.x(t).Cos(2\pi f_c t) \tag{4.7a}$$

$$S(w) = \frac{V_c}{2}[X(w + 2\pi f_c) + X(w - 2\pi f_c)] \tag{4.7b}$$

In (4.7b) the first term is the mirror image of the spectrum shifted to the neighborhood of the carrier frequency. This process is depicted in Figure 4.6.

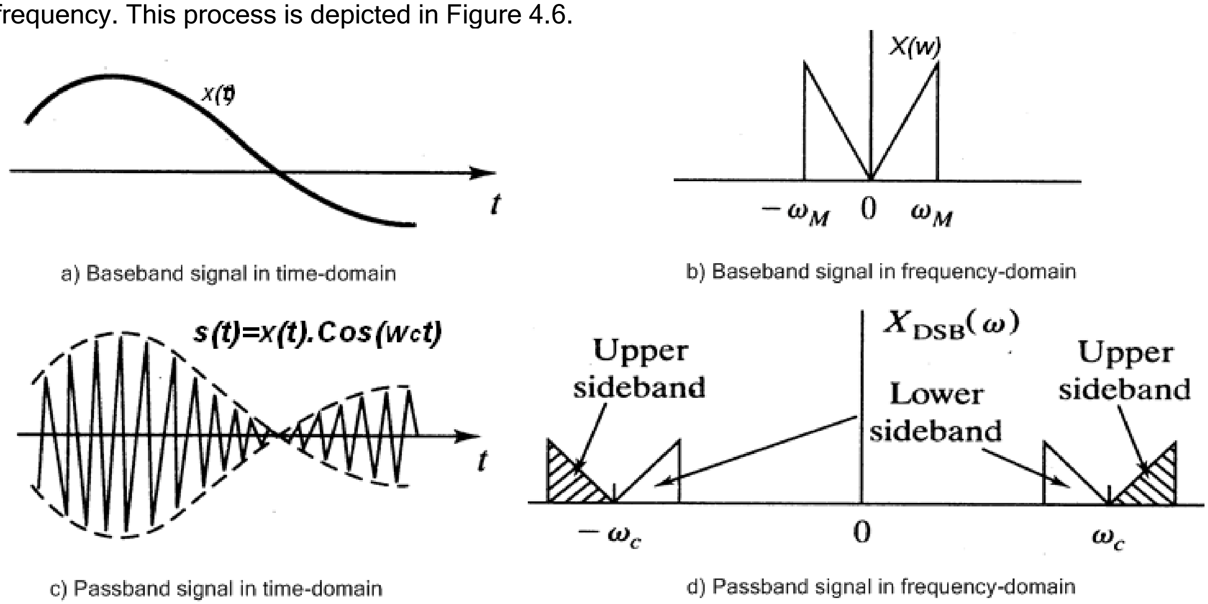


Figure 4.6 Time and frequency-domain illustrations of AM processes.

Observations:

- As it can be seen from Figure 4.6d (d), we have two side bands in each spectra, namely, the "Lower Side Band *(LSB)"* and the "Upper Side Band *(USB)."*
- Having both of them in many applications is waste of bandwidth and only one is used since the information in the other half is just a mirror image. If they are both kept then the technique is called a "Double Side Band Amplitude Modulation *(DSB-AM)"* as show in Figure 4.6 above.
- In the case when there is an explicit frequency component at the carrier frequency $w = w_C$ then this is called "modulation with carrier" including the AM broadcasting systems.
- If there is no explicit carrier frequency then these systems are classified as "Suppressed Carrier *(SC)"* systems, which have smaller power budget, which it will be discussed in detail later. As it can be seen in Figure 4.6 that there is no distinct frequency component at carrier frequency w_c. Therefore, this setup is properly abbreviated by "DBS-SC Amplitude Modulation."

4.2.2 Demodulation Process:

Assuming the transmitted signal did not experience any degradation in the channel, we can mix the incoming signal with an identical copy of the carrier signal:

$$e(t) = x(t)Cos(w_c t).Cos(w_c t) = 1/2.[x(t) + x(t).Cos(2w_c t)] \tag{4.8a}$$

which contains a scaled version of the information signal $x(t)$ and another modulated version at twice the carrier frequency: $w = 2w_c$.

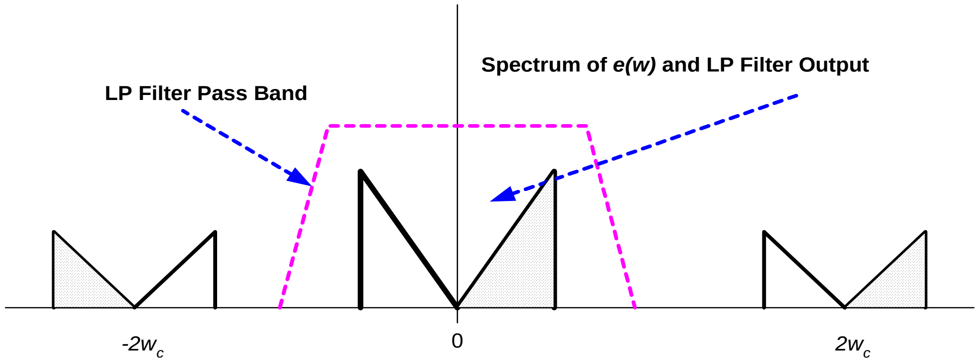
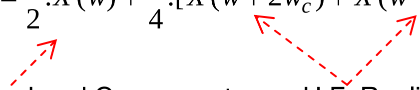


Figure 4.7 Demodulation operation in frequency-domain.

The equivalent operation in the frequency-domain is shown in Figure 4.7 and it is written in the form:

$$E(w) = \frac{1}{2}.X(w) + \frac{1}{4}.[X(w+2w_c) + X(w-2w_c)] \quad (4.8b)$$



If we use a low-pass filter with an appropriate bandwidth the two high-frequency components can be eliminated to yield a replica of the original signal except a scaling factor of 2.0, which can be adjusted by a baseband amplifier to obtain: $y(t) = x(t)$

Observations:

- As it can be seen from the process that the exact recovery needs the channel noise to be small in order this to work.
- Bandwidth of the filter and the linearity of the amplifier are two other factors. However, the most important issue is:

  One needs an exact copy of the local oscillator to generate the carrier signal identical to that of the transmitter. This requires what is commonly known in the literature as "coherent detection" or a reliable carrier recovery system. Thanks to the development in the VLSI circuitry and the availability of low-cost phase-lock loop devices, these are no longer impediments of the AM systems.

Example 4.1: Tone Modulation. Let us assume that the input to the sinusoidal modulator with DSB be another sinusoid:

$$m(t) = V_m.Cos(2\pi f_m t) = V_m.Cos(w_m t) \quad (4.9)$$

The modulator output becomes:

$$\begin{aligned} v_o(t) &= V_m V_c.Cos(w_m t)Cos(w_c t) \\ &= V_m V_c / 2.[Cos((w_c + w_m)t) + Cos((w_c - w_m)t)] \end{aligned} \quad (4.10)$$

The spectrum of this signal is obtained from the Fourier transform of these two sinusoids:

$$\begin{aligned} V_o(w) = A\{&\delta[w-(w_c - w_m)] + \delta[w+(w_c - w_m)] \\ &+ \delta[w-(w_c + w_m)] + \delta[w+(w_c + w_m)]\} \end{aligned} \quad (4.11)$$

where the magnitude of spectral impulses is: $A = \pi V_m V_C / 2$. As it is clear from the Figure 4.8, the resultant spectrum is composed of two impulses at locations $\pm w_m$ away from the carrier frequency. As expected, there are two additional impulses at mirror image frequencies as depicted in Figure 4.8.

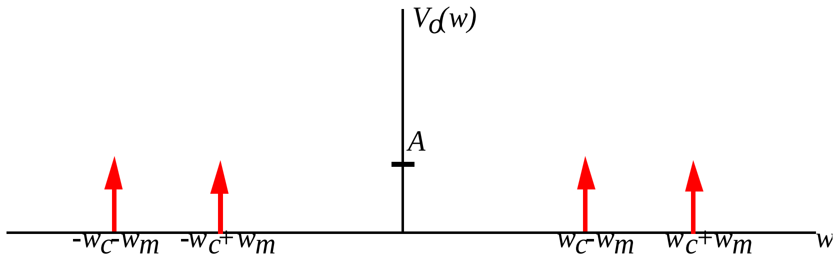


Figure 4.8 Tone modulation (single sinusoid) in frequency-domain.

Observations:

- There is no information at the carrier frequency. Therefore, this is an example for Suppressed Carrier modulation (DSB-SC) case.
- LSB and USB do contain the same information. Hence, it is redundant to transmit both. There are many systems where only one is transmitted and those modulation techniques are called “Single Side Band (SSB)” modulation.

4.3 Double Side Band AM with Carrier Information (Ordinary AM)

Previously, the transmitted information did not contain carrier frequency explicitly; the carrier information at the receiver had to be extracted from the incoming signal coherently using a timing recovery sub-system. This makes the receiver design costlier in order to have good quality reception.

Alternately, one can include an explicit carrier information term in the signal. This would lower the receiver design cost at the expense of an increased power budget; that is, the additional power is needed for the transmission of the carrier. Modulation techniques of this type are called “asynchronous or non-coherent modulation” techniques. However, in the case of AM transmission, it is commonly known as the “ordinary AM” or more precisely the “DSB-AM with Carrier”. Typical applications include broadcasting speech, music (audio) and TV (video) in the frequency band 530-1,600 KHz.

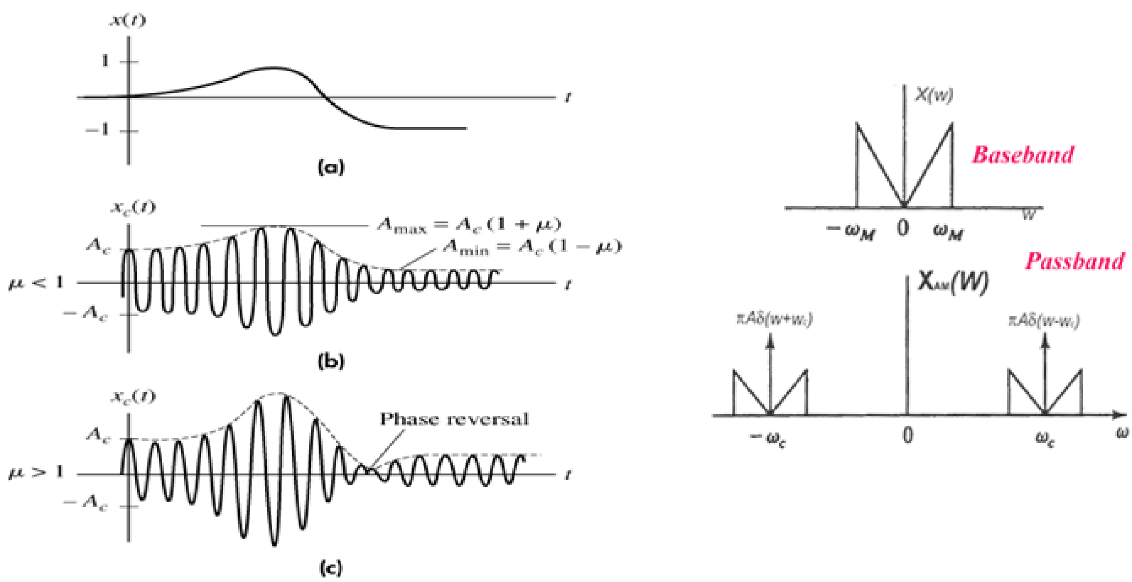


Figure 4.9 Left: ordinary AM (with carrier) signals before and after modulation (Carlson p.153) and right: AM spectra before (baseband) and after modulation on the right (passband).

Time-domain operation in this case is given by:

$$x_{AM}(t) = A.Cos(w_c t) + x(t).Cos(w_c t) = [A + x(t)].Cos(w_c t) \qquad (4.12)$$

Let us again take the Fourier transform to see the impact of the first term in spectra:

$$X_{AM}(w) = (1/2).[X(w+w_c) + X(w-w_c)] + \pi A.[\delta(w+w_c) + \delta(w-w_c)] \quad (4.13)$$

The last two terms are delta functions located at: $\pm w_c$ and represent the carrier information. These delta functions can be clearly seen from Figure 4.9.

The recovery of the original signal at the receiver side is done by low-cost envelope detectors, which will be discussed later in this chapter. In order to have the envelope detectors work efficiently, we must have:

$$A + x(t) > 0 \quad or \quad A \geq -x(t)\big|_{\min} \quad for\ all\ t. \quad (4.14)$$

This is more commonly expressed in terms of a quantity known as the AM modulation index:

$$\mu \equiv \frac{-x(t)\big|_{\min}}{A} \leq 1 \quad (4.15)$$

The case for recoverable AM ($\mu < 1$) case is shown in Figure 9.b. However, if $\mu > 1$ then we have the case of "Over Modulation," which is shown in Figure 9.c. For the case of over-modulated AM waveform, most envelope detectors and a number of other demodulators cannot be used in the design of such an AM receiver. Only, the synchronous detector class is applicable, which pays attention the phase reversal issue But they are comparatively more expensive with respect to asynchronous detection.

In communication systems performance, the percentages of power budgets for sidebands and the carrier are very critical since the power spent on sidebands are used for transmitting the information bearing signal. Let us consider the DSB-AM signal:

$$x_{AM}(t) = A.Cos(w_c t) + x(t).Cos(w_c t) \quad (4.16)$$

where the first term is the carrier information and the last one is the signal. The power in each of these is, respectively:

$$P_C = \frac{1}{T}\int_0^T A^2.Cos^2(w_C t)dt = \frac{1}{2}.A^2 \quad (4.17)$$

$$P_S = \frac{1}{T}\int_0^T x(t)^2.Cos^2(w_C t)dt = 1/2.\overline{x^2(t)} \quad (4.18)$$

The total power is simply the sum of these two terms:

$$P_{Total} = P_C + P_S = \frac{1}{2}[A^2 + \overline{x^2(t)}] \quad (4.19)$$

Sideband power percentage is defined as the ratio of the sideband power to the total power:

$$Sideband\ Power\ Percentage \equiv 100.\frac{P_S}{P_{Total}} = \frac{\overline{x^2(t)}}{A^2 + \overline{x^2(t)}} 100\% \equiv \eta \quad (4.20)$$

Example 4.2: Suppose that the modulation indices are $\mu = 0.3, 0.5\ and\ 1.0$ and a sinusoidal tone input is modulated: $x(t) = \alpha.Cos(w_x t)$. Let us compute the ordinary AM waveform $x_{AM}(t)$ and the corresponding sideband power percentages for these three cases.

$$\mu = \frac{-x(t)\big|_{\min}}{A} = \frac{\alpha}{A}$$

$$x(t) = \alpha.Cos(w_x t) = \mu A.Cos(w_x t)$$

$$x_{AM}(t) = [A + x(t)].Cos(w_c t) = A.[1 + \mu.Cos(w_x t)].Cos(w_c t)$$

Let us use the value $\overline{x^2(t)} = (\mu A)^2 / 2$ and the result is given by:

$$\eta = \frac{(\mu A)^2/2}{A^2 + (\mu A)^2/2} 100\% = \frac{\mu^2}{2+\mu^2} 100\% = \begin{cases} \frac{1}{2+1} 100\% = 33\% & if \;\; \mu = 1.0 \\ \frac{0.25}{2+0.25} 100\% = 11.11\% & if \;\; \mu = 0.5 \\ \frac{0.09}{2+0.09} 100\% = 4.3\% & if \;\; \mu = 0.3 \end{cases}$$

Observation: At best, only 33% of the total power budget goes to the transmission of information and 67% is spent on the carrier signal. Hence, the ordinary AM does not have a favorable power budget.

4.4 Modulators

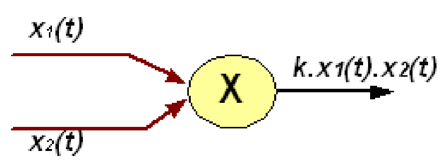
There are several techniques for achieving DSB-SC modulation:

- Analog Multipliers.
- Chopper modulators.
- Pulse modulators.
- Square-law and other non-linear devices, such as IC multipliers.
- Direct-tuned circuit modulators.

Example 4.3: Multiplier modulator is achieved by realizing the multiplication of two signals, namely, the signal and the carrier information as shown in Figure 4.10 :

$$k.x_1(t).x_2(t)$$

The multiplication operation cannot be realized easily. Either non-linear characteristics of active devices are used or alternatively, they can be implemented by using logarithmic and exponential amplifiers, where addition operation replaces to replace multiplication-- and OPAMP circuits.



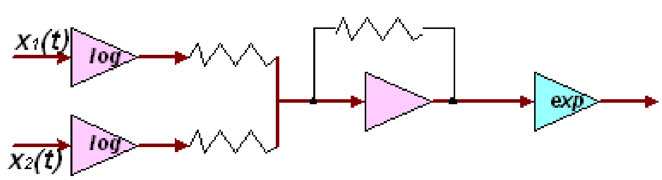


Figure 4.10 Analog Multiplier Modulator using log/exp amplifiers.

Example 4.4: Pair of Balanced Square-Law Devices, such as a linear FET Modulator.

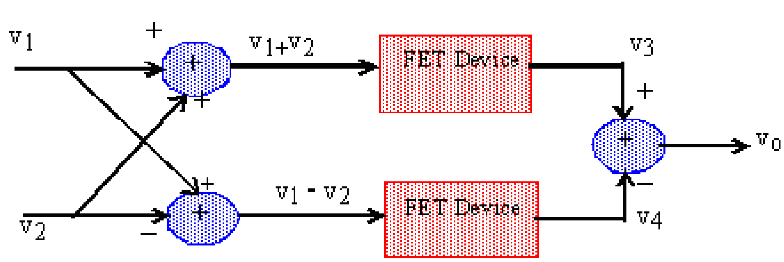


Figure 4.11 Balanced FET Pair Modulator

If we assume that a properly biased and linearly amplified FET device in Figure 4.11 is a perfect square-law device then we can write:

$$v_3 = K_s(v_1^2 + v_2^2 + 2v_1v_2) \quad (4.21a)$$

$$v_4 = K_s(v_1^2 + v_2^2 - 2v_1v_2) \quad (4.21b)$$

and the output is the algebraic sum of the two:

$$vo = v_3 - v_4 = 4.K_s v_1.v_2 \quad (4.22)$$

In practice, one of the inputs is the message signal to be transmitted and the other one is the carrier signal. A low-cost IC device from Analog Devices, Inc. has been depicted in Figure 4.12.

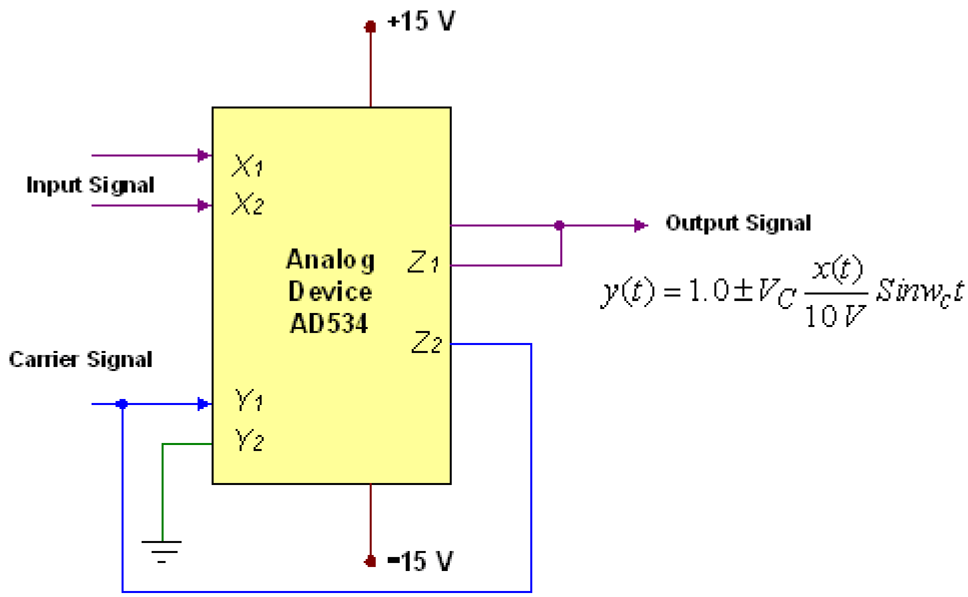


Figure 4.12 A VLSI Multiplier Modulator from Analog Devices (Courtesy of Analog Devices, Inc.)

In addition to these there are few other modulators including *Chopper Modulator* and *Non-linear Modulators* using balanced diodes . However, one of the most frequently used modulator circuits are called Switching Modulators and Ring Modulators based on four identical diodes either in a bridge or ring topology as shown in Figure 4.13. Their common identifying representation is the use of a switching function modeled *k(t)* by a square pulse-train.

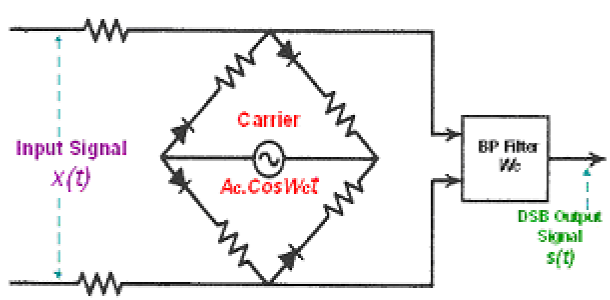


Figure 4.13 Chopper (Switching) Modulator

$$k(t)=\frac{1}{2}+\frac{2}{\pi}.\sum_{n=1,3,5,\cdots}\frac{(-1)^{(n-1)/2}}{n}Cos(nw_c t)$$
$$=\frac{1}{2}+\frac{2}{\pi}Cos(w_c t)-\frac{2}{3\pi}Cos(3w_c t)+\frac{2}{5\pi}Cos(5w_c t)+\cdots \quad (4.23)$$

It is easy to see that we have all the odd harmonics of the fundamental frequency term in addition to the ½ factor of the DC bias. When we multiply this pulse-train expansion with the input signal we get:

- baseband copy scaled by a factor of ½ ,
- DSB modulated passband at w_c and
- Infinitely many harmonics at nw_c, where n is any odd integer.

n compact form:

$$v(t)=x(t).k(t)=\frac{1}{2}x(t)+\frac{2}{\pi}x(t).Cos(w_c t)-\frac{2}{3\pi}x(t).Cos(3w_c t)+\cdots \quad (4.24a)$$

Let us take the Fourier transform of both sides of either (4.24a):

$$V(w)=\frac{1}{2}X(w)+\frac{1}{\pi}\sum_{n=1,3,5,\cdots}\frac{(-1)^{\frac{n-1}{2}}}{n}[X(w+nw_c)+X(w-nw_c)] \quad (4.24b)$$

Using a properly designed bandpass filter with a center frequency at $w=w_c$ we can achieve the following:

- Eliminate the baseband copy of the spectrum (first term),
- Keep the bandpass modulated term, i.e. *n=1* case, and
- Eliminate the terms at multiples of the carrier frequency other than *n=1*.

This will result in the desired suppressed carrier DSB-AM output.

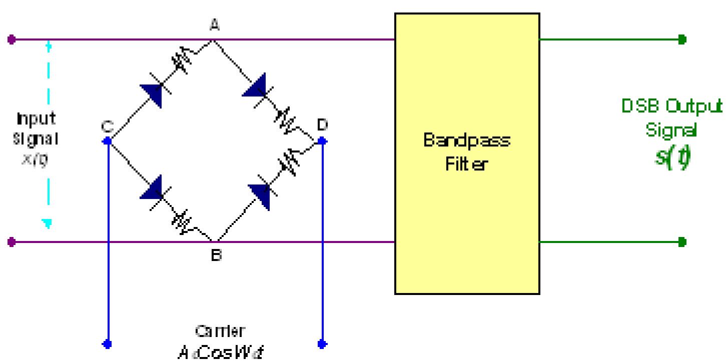


Figure 4.14 Chopper (Switching) modulator using a diode bridge

Example 4.5: Let us analyze the shunt and series bridge circuit in Figure 4.14 with identical four diodes.

1. In the half cycle when point C is more positive than point D
   - All diodes are conducting.
   - Point A and B have the same voltage and
   - The input to BPF is shorted, which results with an output of zero.
2. In the next half cycle, D is more positive this time then:
   - All diodes are open.
   - *x(t)* shows up at the input of the BPF.

Example 4.6: Frequency Converter (Mixer) as an AM modulator.

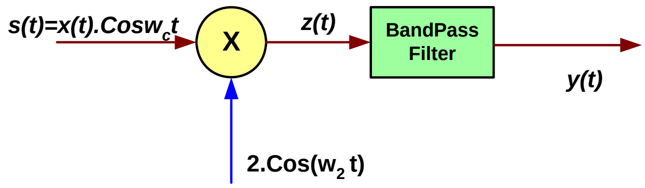


Figure 4.15 Frequency Converter (Mixer)

The output of the multiplier at the above frequency converter is the product of both inputs:

$$z(t) = 2x(t)Cosw_c t.Cosw_2 t$$
$$= x(t)[Cos((w_c - w_2)t) + Cos(w_c + w_2)t]$$

If we choose $w_2 = w_c - w_{IF}$, where w_{IF} is commercially known as the Intermediate frequency (IF), which set at 455 KHz for AM radio transmission by the International Telecommunication Union (ITU), we can write:

$$z(t) = x(t)[Cosw_{IF}t + Cos(2wc - w_{IF})t]$$

On the other hand, we choose $w_2 = w_c + w_{IF}$, the result becomes:

$$z(t) = x(t)[Cosw_{IF}t + Cos(2w_c + w_{IF})t]$$

In either case, a bandpass filter at the output will pass the first term $x(t)Cosw_{IF}t$ but the second higher frequency term will be suppressed to yield the desired output: $x(t)Cosw_{IF}t$. This frequency conversion to IF frequency for modulation purposes is commonly known as the "heterodyning" in the literature. The signal at IF frequency is later up-converted to the frequency band of the transmitting station, which is unique for each and every radio station. The radio receiver has a similar stage called down-conversion to bring the signal back to the IF range before it is completely demodulated to the baseband.

Example 4.7: Consider the circuit given in Figure 4.16 where the signal and the carrier are connected in series (added). Generation of ordinary AM using a single diode as switch.

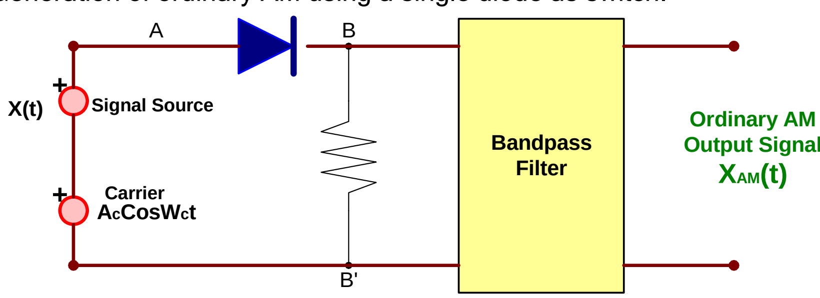


Figure 4.16 Ordinary AM (DSB AM with carrier) generation using a single diode.

The signal at point A is the sum of the signal plus the carrier:

$$x_A(t) = x(t) + A_c.Cosw_c t$$

and the signal at section BB' is the switched version as discussed earlier:

$$x_{BB'}(t) = [x(t) + A_c.Cosw_c t].k(t)$$

When we expand the switch into its Fourier Cosine series as we did earlier and multiply with the signal at point A we have the switched modulated signal:

$$x_{BB'}(t) = [x(t) + A_c.Cosw_c t].[\frac{1}{2} + \frac{2}{\pi}Cosw_c t - \frac{2}{3\pi}Cos3w_c t + \frac{2}{5\pi}Cos5w_c t + \cdots]$$
$$= \frac{A_c}{2}Cosw_c t + \frac{2}{\pi}.x(t).Cosw_c t + Baseband\ Terms + Higher\ Order\ Terms \quad (4.25)$$

The first two terms in (4.25) make up the DSB-AM signal, whereas the last two are the unwanted baseband term and infinitely many odd-order high frequency terms. These unwanted signals are suppressed by an appropriately designed BPF with a center frequency at $w = w_C$ and a bandwidth twice the input signal bandwidth.

4.5 Demodulation of AM Signals

Demodulation of all AM signals, which includes SC-AM, SSB-AM and DSB-AM, is normally achieved in two stages. As shown in the block diagram in Figure 4.17, the received signal *r(t)* is first demodulated to an intermediate frequency (IF). Here the first stage of filtering is applied to eliminate RF level unwanted signals and then a common reference carrier w_c is applied to the mixer to down convert the information-bearing signal. Finally, an appropriate LP filter extracts the baseband information.

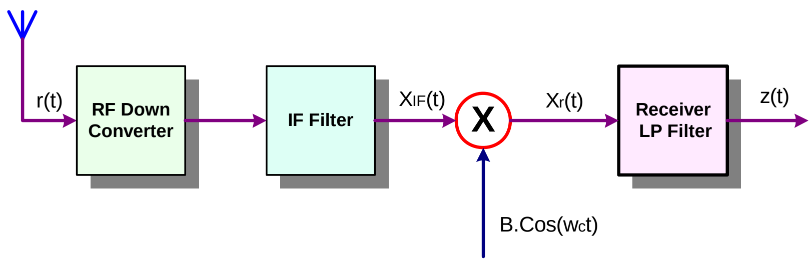


Figure 4.17 Block Diagram of an AM Demodulator with an Intermediate Frequency Stage.

$$x_{IF}(t) = \begin{cases} x(t).Cos(w_c t) & \begin{cases} for\ DSB-AM\ \ with\ x(t) \geq 0 \\ for\ SC-AM\ \ with\ \overline{x(t)} = 0 \end{cases} \\ & OR \\ \frac{x(t)}{2}.Cos(w_c t) \pm \frac{\hat{x}(t)}{2}.Sin(w_c t) & for\quad SSB \end{cases} \quad (4.26)$$

Let us recall $x(t)$ is the baseband original signal and $\hat{x}(t)$ is the Hilbert transform of the signal needed during the SSB generation.

Observations:

- We can use average envelope detection for DSB-AM case.
- We can use peak envelope detection for DSB-AM case.
- We must use synchronous detection for SC-AM and SSB to be studied later. However, we can also use this type of detection for DSB-AM as well.

Rectifier Detector as an Envelope Detector:

We have discussed earlier the usage of a rectifier switch function $k(t)$ employed for extracting a replica of the transmitted signal. Let us revisit the problem as an envelope detector, which has been the workhorse since the early days of radio transmission and the broadcasting of messages. A detector based on a simple diode followed by an RC filter network is shown in Figure 4.18.

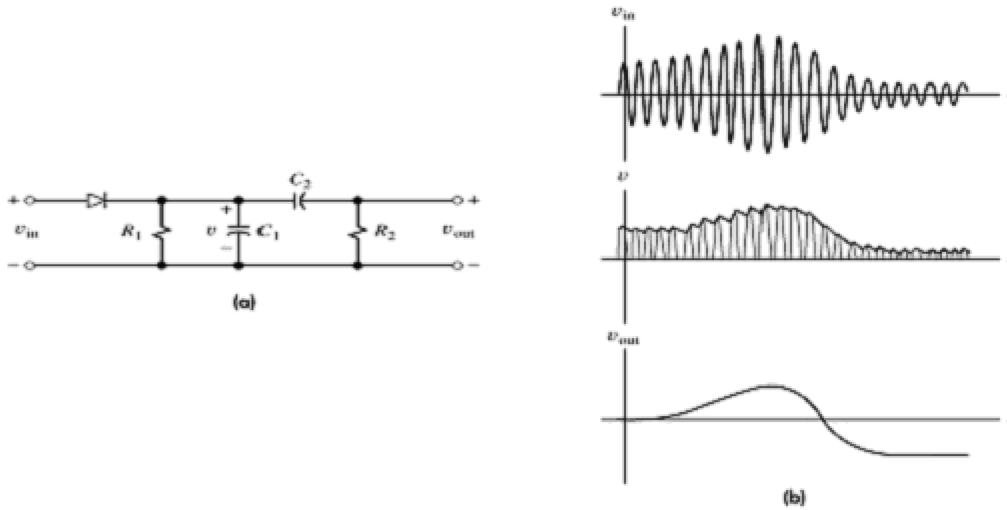


Figure 4.18 AM Demodulation Using Envelope Detector. (Carlson p. 177)

Envelope detector attempts to track the transmitted signal closely (desired). However, when the time constant τ of the network is too large and the detector misses the peaks. On the other hand, when the time constant τ is too small and the detector output dives deep and this results in large ripples.

Assuming the detector is able to track the incoming signal, the rectified signal at the input of the low-pass filter will:

$$z(t) = \{[A + x(t)]Cosw_c t\}.S(t) = \frac{1}{\pi}[A + x(t)] + hf \ terms\ of\ order\ N = 1,2,3,\cdots \quad (4.27)$$

High frequency harmonic terms are suppressed by the usual low-pass filter and the D.C (constant) term is blocked by the capacitor in the circuit to yield in a close replica of the original signal $x(t)$.

In order to closely track the signal, we must have a value for τ in the range of:

$$1/w_c << \tau << 1/2\pi B \quad (4.28)$$

where B is the bandwidth of $x(t)$. The output will have an exponential decaying form:

$$v_o(t) = E(t).e^{-t/\tau} \quad (4.29a)$$

Since the left inequality is satisfied then we can approximate the exponential form with a line to have:

$$v_o(t) \approx E(t).(1 - t/\tau) \quad (4.29b)$$

It is worth noting that the slope of discharge is simply: $-E(t)/\tau$. For capacitors to follow $E(t)$ closely, we must have:

$$|dv_o(t)/dt| = E(t)/\tau \geq |dE(t)/dt| \quad (4.30a)$$

For the tone modulation discussed earlier, this constraint becomes:

$$\tau \leq \frac{1}{w_m}\left(\frac{\sqrt{1-\mu^2}}{\mu}\right) \quad (4.31)$$

In the case of suppressed carrier DSB signal, we need to provide a sync signal to fine tune the oscillator in the receiver as shown in Figure 4.19.

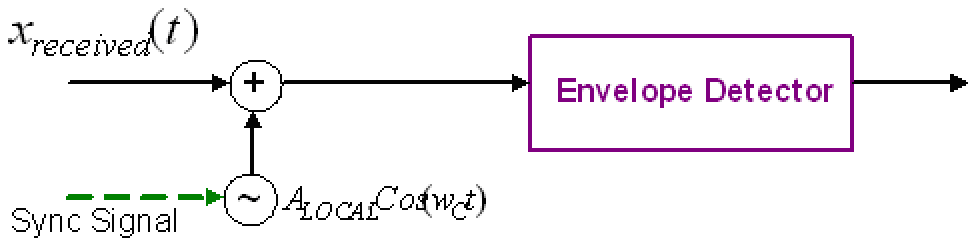


Figure 4.19 Envelope detection (coherent) for suppressed-carrier modulation.

Here, a local crystal oscillator generates a generic frequency, but it must be tuned to the precise carrier frequency w_C from a carrier recovery circuit, which is normally a phase-lock loop network.

4.6 Quadrature Amplitude Modulation (QAM)

We have observed in the previous sections that DSB spectra occupy twice the bandwidth required of their baseband correspondents. This disadvantage can be mitigated by transmitting two different DSB signals using the same carrier in a in-phase and quadrature forms (orthogonal forms) in many new and emerging telecommunication applications. These systems are classified as quadrature modulation schemes since they contain quadrature component. If the modulated components of the baseband signals are the amplitudes, then the systems achieve that are called "Quadrature Amplitude Modulation (QAM) techniques. There are many recent applications of QAM including the color transmitter in the NTSC-TV, the signal placement in V.90 and V.92 Modem standards and the high-rate date communication using the family of DSL products.

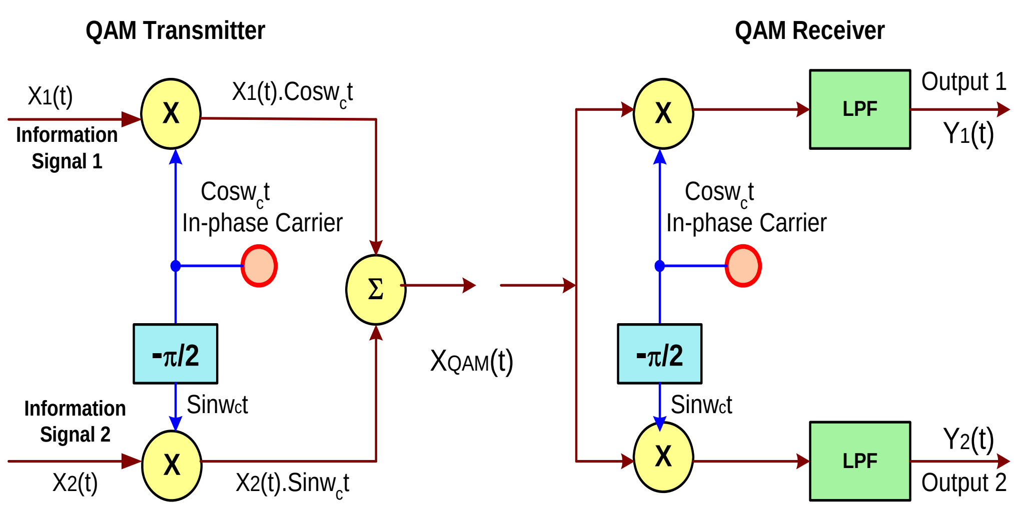


Figure 4.20 Block Diagram for a QAM system with Transmitter and Receiver sections.

In this setup, multipliers and -90^o phase-shifters are used for converting the signal into passband and shifting the phase of a sinusoidal carrier signal by $-\pi/2.$, respectively. As we can clearly see from the transmitter blocks in Figure 4.20, the transmitted QAM signal is the sum of two DSB-modulated components coming from upper and lower paths:

$$x_{QAM}(t) = A_c.x_1(t).Cosw_c t + A_c.x_2(t).Sinw_c t \qquad (4.32)$$

Next we try to demonstrate that exact recovery of these two information signals $\{x_1(t), x_2(t)\}$ is possible. At the receiver side the reconstruction of the upper baseband signal, i.e., $x_1(t)$, is simply:

$$x_1(t) = x_{QAM}(t).A'_c.Cosw_ct$$
$$= A'_c.A_c[x_1(t)Cosw_ct + x_2(t).Sinw_ct].Cosw_ct \quad (4.33a)$$
$$= \frac{1}{2}A'_c.A_c.x_1(t) + \frac{1}{2}A'_c.A_c.x_1(t).Cos2w_ct + \frac{1}{2}A'_c.A_c.x_2(t).Sin2w_ct$$

and the output from this branch is:

$$y_1(t) = \frac{1}{2}A'_c.A_c.x_1(t) = K.x_1(t) \quad (4.33b)$$

The first term is the desired base-band signal for the upper information signal since it is directly proportional, with an amplification factor K, to the original input in the QAM transmitter. The second and third terms are band-pass signals in the neighborhood of $2w_c$ and they are completely discarded by the LP filter in the upper path of the QAM receiver. Similarly, the output from the lower path of the receiver will produce:

$$y_2(t) = \frac{1}{2}A'_c.A_c.x_2(t) = K.x_2(t) \quad (4.34)$$

Therefore, we have completely recovered signals from both channels without distortion.

On the other hand, if the overall setup is not perfectly synchronized, the system will have co-channel interference, i.e., leakage from one signal to the other. To be more specific on this let us assume that the local carrier at the demodulator has a phase jitter $A'_cCos(w_ct + \theta)$, in this case, the operations in the upper path will yield:

$$z_1(t) = A'_c.[x_1(t).A_cCosw_ct + x_2(t).A_cSinw_ct].Cos(w_ct + \theta)$$

Using trigonometric identities we expand this expression:

$$z_1(t) = \frac{1}{2}.A'_c.A_c.x_1(t).Cos\theta + \frac{1}{2}.A'_c.A_c.x_1(t).Cos(2w_ct + \theta)$$
$$-\frac{1}{2}.A'_c.A_c.x_2(t).Sin\theta + \frac{1}{2}.A'_c.A_c.x_2(t).Sin(2w_ct + \theta) \quad (4.35a)$$

The second and fourth terms will be filtered by the upper LP filter and they will not effect the recovery of the original input. Let us now write the remaining terms at the output of this upper filter:

$$y_1(t) = \frac{1}{2}.A'_c.A_c.x_1(t).Cos\theta - \frac{1}{2}.A'_c.A_c.x_2(t).Sin\theta \quad (4.35b)$$

If we study this last equation term by term we see that the first term is proportional to the signal transmitted through the upper branch of the QAM transmitter. If $\theta \neq 90^o$ then we will recover it at the output of the upper LP filter, which is good. But, the second term is the interference from the lower path, i.e., the quadrature term. Since that term is a base-band signal, just as the first term, it will corrupt the output $y_1(t)$. Depending upon the actual value of the multiplicative factor $Sin\theta$, the legitimate signal $x_1(t)$ might even be masked completely. This results in a total jam. It is not difficult to guess that this is the way smart jammers work in the real-life.

4.7 Single Side Band Amplitude Modulation (SSB) and VSB

In many applications, the spectrum of the input signal has even symmetry, such as speech, audio, and imagery. That is, the spectrum in one half contains all the information and the other half is completely redundant. If this redundancy is removed there will be no information loss and the required channel bandwidth will be halved. Systems using only one side band are called “Single Side-Band

(SSB) Modulation Systems". However, this reduction in bandwidth does not come free. The receiver needs to have a synchronous detector as in the case of DSB-SC modulation. In addition, there are two SSB scenarios:

- Lower Side-Band SSB (LSB) and
- Upper Side-Band SSB (USB).

In the first one, only the frequency bands to the left of the carrier are preserved and the higher frequency components are suppressed. Similarly, the opposite is valid for the USB case.

4.7.1 Analysis of SSB Modulation: The modulation and demodulation operations for SSB systems look very simple; however, practical implementation is a major design challenge for two reasons: First, the modulator calls for an ideal band-pass filter; and second, the demodulator requires a synchronous carrier.

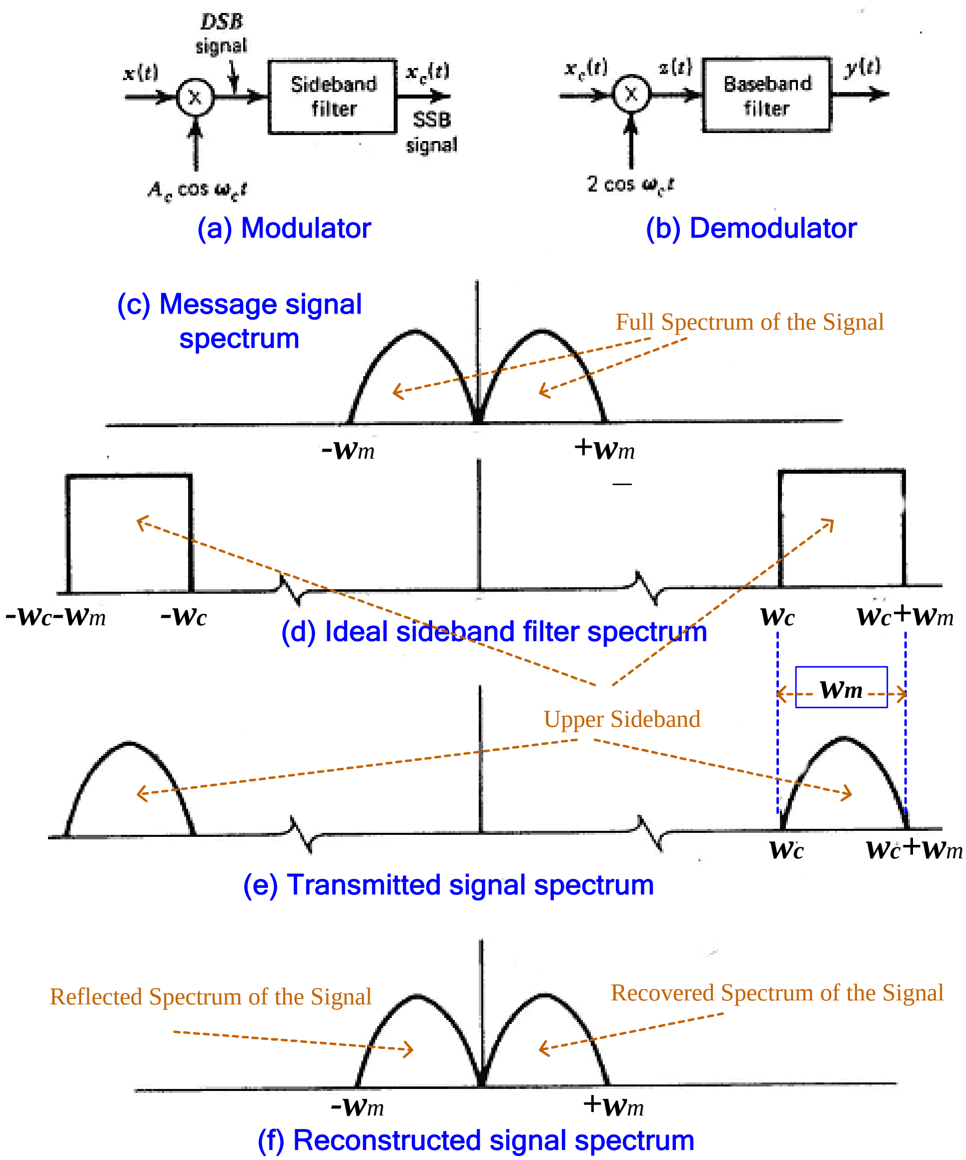
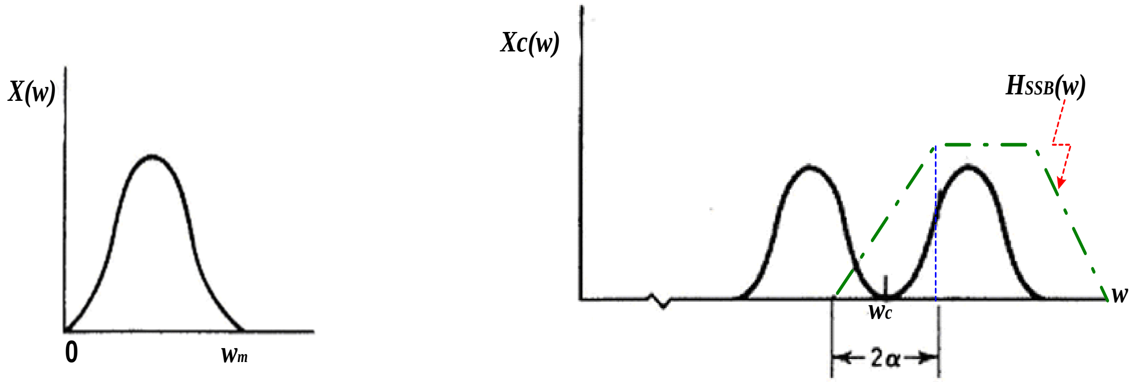


Figure 4.21. Single Side Band (SSB) modulation. a.) modulator, b.) demodulator, c.) signal spectrum, d.) ideal side-band filter, e.) transmitted signal spectrum, and f.) reconstructed signal spectrum.

In practice, vestigial side band (VSB) filters of the next section is used to defray the problem. The characteristics of one such filter in frequency domain is shown in Figure 4.22.



a.) Information Signal Spectrum b.) SSB Filter Superimposed on DBS Spectrum

Figure 4.22. Single Side Band (SSB) modulation using VSB filters instead of ideal BP filters.

4.7.2. Hilbert Transform: There are some reasonably complex VSB filters used in practice. But the SSB is more readily generated by a phase-shift or Hilbert transform method as depicted in Figure 4.23.

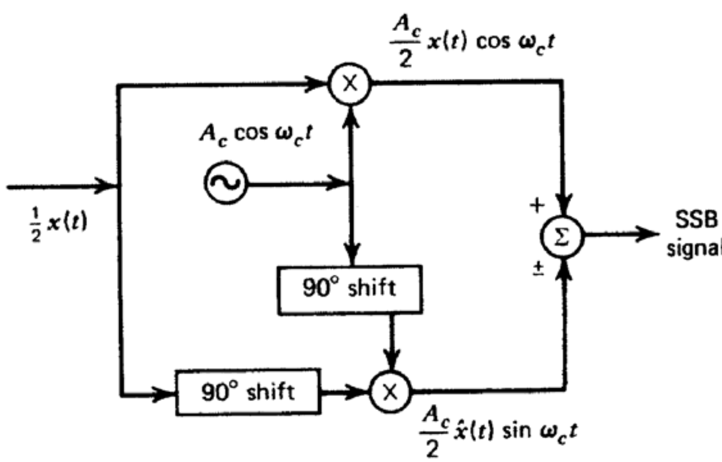


Figure 4.23. Single Side Band (SSB) generation using Hilbert transform.

To analyze this technique let us assume that the information signal can be represented by a finite number of Fourier terms:

$$x(t) = \sum_{i=1}^{n} X_i Cos(w_i t + \theta_i) \qquad where \quad f_n \le f_m \tag{4.36}$$

Then the upper side-band SSB signal, (similar equation can also be written for lower side band, as well) corresponding to $x(t)$ is given by:

$$x_c(t) = \frac{A_c}{2} \sum_{i=1}^{n} X_i Cos[(w_c + w_i)t + \theta_i] \tag{4.37}$$

From Fourier theory the last expression can be expanded into cosine and sine series:

$$x_c(t) = \frac{A_c}{2}\{[\sum_{i=1}^{n} X_i Cos(w_i t + \theta_i)]Cosw_c t - [\sum_{i=1}^{n} X_i Sin(w_i t + \theta_i)]Sinw_c t \qquad (4.38a)$$

However, it is customary to write this series in a more compact form:

$$x_c(t) = \frac{A_c}{2}.x(t).Cosw_c t - \frac{A_c}{2}.\hat{x}(t).Sinw_c t \qquad (4.38b)$$

where $\hat{x}(t)$ is known as the Hilbert transform of $x(t)$, which is defined by:

$$\hat{x}(t) = \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{x(\alpha)}{t-\alpha}d\alpha \qquad (4.39)$$

but it is also easily written in terms of Fourier coefficients from (4.38a):

$$\hat{x}(t) = \sum_{i=1}^{n} X_i Sin(w_i t + \theta_i] \qquad (4.40)$$

Since this process a transform operation it can be configured as the response of a Hilbert Filter with a system function:

$$H_{HF}(w) = -j.\text{sgn}(w) = \begin{cases} -j = e^{-j\pi/2} & if \quad w > 0 \\ j = e^{+j\pi/2} & if \quad w < 0 \end{cases} \qquad (4.41)$$

Graphically this can be shown as in the following set up.

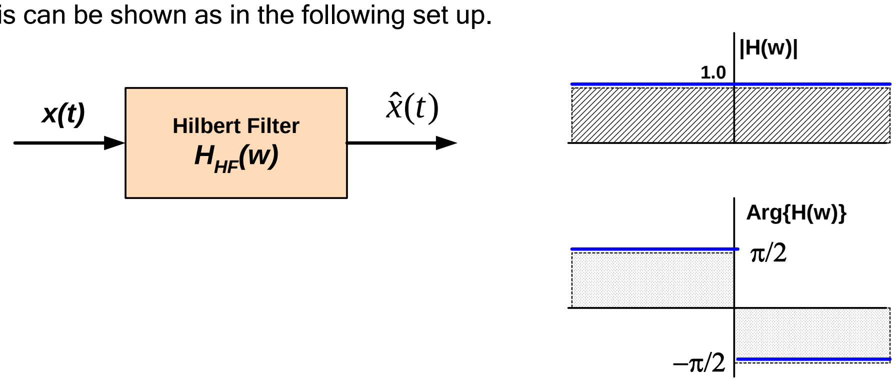


Figure 4.24. System representation of Hilbert transform and its amplitude and phase responses.

Using this filter set-up we have can eliminate one of the difficulties mentioned above. However, the synchronous carrier issue needs to be dealt with as well. In order to handle that, it is common practice to add a low power pilot tone as the carrier information at the transmitter, which will be evident in the discussion of demodulation of SSB systems below.

4.7.3 Demodulation of SSB Signals:

There are two approaches to demodulate single side-band suppressed carrier (SSB-SC) signals to recover the original DSB information:

- Coherent (synchronous detection)
- Envelope detector for asynchronous detection.

Coherent (synchronous) demodulation:

In this approach, we assume that the exact carrier frequency and phase information is locally available to the receiver, where the received signal

$$x_{SSB}(t) = x(t).Cos(w_c t) \pm \hat{x}(t).Sin(w_c t) \qquad (4.42)$$

is multiplied by the identical carrier signal:

$$x_{SSB}(t).Cos(w_c t) = x(t).Cos^2(w_c t) \pm \hat{x}(t).Sin(w_c t).Cos(w_c t)$$
$$= \frac{1}{2}x(t) + \frac{1}{2}x(t).Cos(2w_c t) \pm \frac{1}{2}\hat{x}(t).Sin(2w_c t) \quad (4.44)$$

These two equations represent the transmitter and receiver operations in a SSB system-based on Hilbert transforms, where

- First term is the scaled version of the baseband information.
- Last two terms are modulated signals in the neighborhood of $2w_c$.

As in all of the similar demodulation stages, they are easily suppressed by a low-pass filter. It is worth noting that this scheme is identical to the coherent demodulation of DSB-SC system Therefore, the designs for DSB-SC demodulators can be used without any modification.

Envelope detector for SSB with a carrier (SSB+C) signal:

In this case, there is an explicit carrier component in the received signal:

$$\phi(t) = A.Cos(w_c t) + [x(t).Cos(w_c t) \pm \hat{x}(t).Sin(w_c t)]$$
$$= [A + x(t)].Cos(w_c t) \pm \hat{x}(t).Sin(w_c t) = E(t).Cos(w_c t + \theta) \quad (4.45)$$

where the envelope term $E(t)$ is defined by:

$$E(t) = \{[A + x(t)]^2 + \hat{x}^2(t)\}^{1/2}$$
$$= A.\{1 + \frac{2x(t)}{A} + \frac{x^2(t)}{A^2} + \frac{\hat{x}^2(t)}{A^2}\}^{1/2} \quad (4.46)$$

In the case when $A >> |x(t)|$ the last two terms of the last equation can be ignored to result in:

$$E(t) \approx A[1 + \frac{2x(t)}{A}] = A + 2x(t) \quad (4.47)$$

Thus, it is possible to recover the original input signal $x(t)$ within a scale and D.C. offset. For a large carrier, the envelope of the SSB signal follows the baseband signal within a D.C. offset and the receiver is simply an envelope detector.

4.7.4 Vestigial Side Band Amplitude Modulation (VSB):

Previously we have seen that simple DSB systems have two complete side bands (redundant), whereas SSB had only one band at a cost of increased system complexity and the necessity of coherent demodulation. A compromise between these two extremes is a scheme called Vestigial Side Band (VSB), which is used in the current TV transmission systems. Here the bandwidth is typically 25% greater than the corresponding SSB, yet it is relatively easy to generate.

As it can be seen from Figure 4.25 -- the magnitude has been plotted as a function of the actual frequency in Hz as it is normally done in the industry-- one side band is gradually cut-off. The lost portion of the desired side band is exactly compensated by a bandwidth expansion of the same amount.

A synchronous detector can detect this signal rather easily. However, if the carrier itself or a pilot tone is also transmitted with the VSB information then an envelope detector such as a rectifier detector can reliably recover original signal. To achieve this side band manipulation we use a pair of VSB Shaping Filters on the transmitting and receiving sides

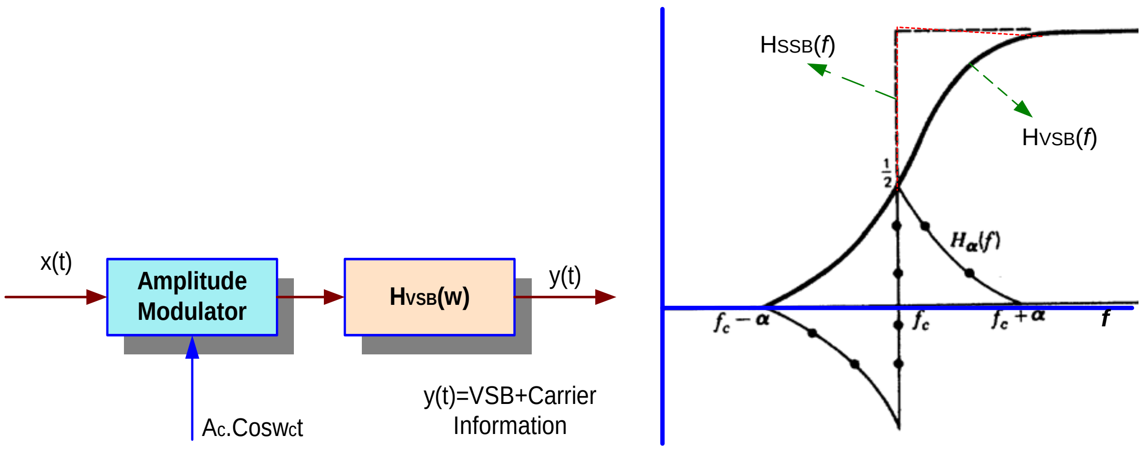
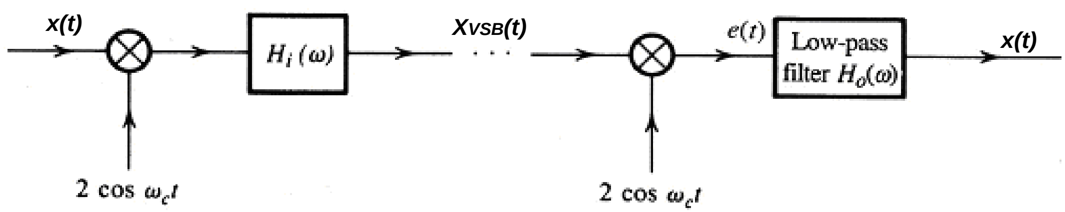


Figure 4.25. Vestigial Side Band (VSB) generation and VSB spectrum as compared to SSB Spectrum.



a) VSB Transmitter **b) VSB Receiver**

Figure 4.26. Vestigial Side Band (VSB) transmitter and receiver functional diagrams.

The current task is to find the characteristics of the pair of shaping filters in Figure 4.26:

$$X_{VSB}(w) = H_i(w).[X(w+w_c) + X(w-w_c)] \tag{4.48}$$

The message signal $x(t)$ must be recoverable from the vestigial signal in time-domain or in frequency domain:

$$x_{VSB}(t) = F^{-1}\{X_{VSB}(w)\}$$

To do that let us multiply $x_{VSB}(t)$ with twice the carrier information to obtain:

$$e(t) = x_{VSB}(t).2.Cos(w_c t) \tag{4.49a}$$

or equivalently in frequency-domain:

$$E(w) = [X_{VSB}(w+w_c) + X_{VSB}(w-w_c)] \tag{4.49b}$$

When we substitute above the expressions for the VSB terms we get:

$$E(w) = [X(w+2w_c) + X(w)]H_i(w+w_c) + [X(w) + X(w-2w_c)].H_i(w-w_c) \tag{4.50}$$

The receiver VSB filter $H_o(w)$ will suppress signal components in the neighborhoods of $\pm 2w_c$ to yield an output signal:

$$Y(w) = X(w).[H_i(w+w_c) + H_i(w-w_c)].H_o(w) \tag{4.51}$$

In order to recover the original input signal *x(t)* without distortion we need to the VSB filter set to satisfy:

$$H_o(w) = \frac{1}{H_i(w+w_c) + H_i(w-w_c)} \quad for \quad |w| \le 2\pi B \tag{4.52}$$

Under this condition, the output is the desired original signal $x(t)$:

$$Y(w) = X(w) \tag{4.53a}$$

and in time-domain:

$$y(t) = x(t) \tag{4.53b}$$

For distortionless transmission, we must have flat spectra for our band of interest $|w| \le 2\pi B$:

$$H_i(w + w_c) + H_i(w - w_c) = 1 \quad for \quad |w| \le 2\pi B \tag{4.52}$$

This is commonly known as the ***Vestigial Condition*** in the business. There are many filters satisfying this condition. In order this filter to be of any use it must be real and physically realizable.

4.7.5 Superheterodyne Receiver:

Transmitters for audio signals used in broadcasting have a two-stage reception structure. The first stage is called the rf-stage and it is tunable and the second stage --if-stage-- is a fixed design common to all receivers. This architecture is known as the superheterodyne receiver. Local oscillator frequency in superheterodyne receivers is the sum of the carrier and IF frequency of 455 KHz.

$$f_{Lo} = f_c + f_{IF} = f_c + 455\ KHz \tag{4.53}$$

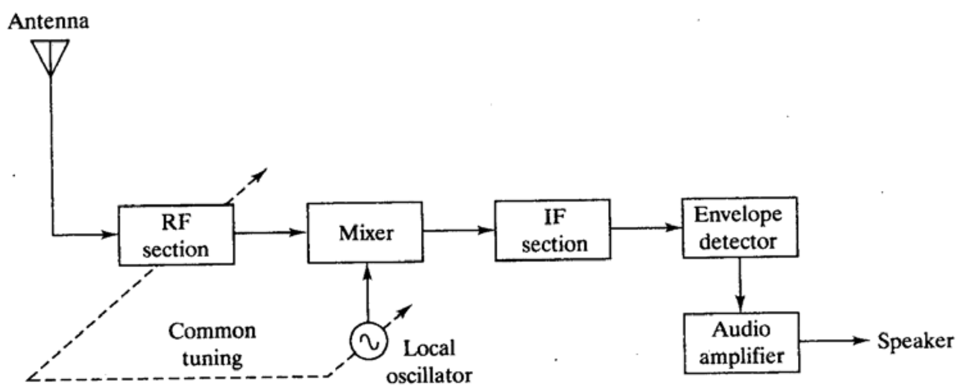


Figure 4.27. Superheterodyne Receiver functional diagram.

Note that the AM carrier frequency range of $530\ KHz \le f_c \le 1600\ KHz$ results in a frequency range of $985\ KHz \le f_c \le 2055\ KHz$ in the up-conversion step, which may lead to *phantom station* tunings. Consider the following radio transmission, $f_c = 1{,}000\ KHz_c$ which needs a local oscillator frequency of $f_{Lo} = 1{,}455\ KHz$. But there is another phantom station out there at $f_c^{'} = 1455 + 455 = 1{,}910\ KHz$. It is not difficult to see that this station will have the same local frequency due to $f_c^{'} - f_{Lo} = 455\ KHz$. In other words the two stations at 1000 KHz and 1910 KHz are mirror sites of each other.

4.8 NTSC TV System

The TV system is used transmit a two-dimensional image with motion, so, it needs a description with 2 spatial variables and time. There are several type of TV systems including NTSC, PAL and SECAM of CCIR and now ATSC or as it is commonly known HDTV. Basic features of TV Standards are tabulated in Table 4.3.

Table 4.3 TV System Parameters

System Parameters	NTSC	CCIR	HDTV
Aspect Ratio (H/V)	4/3	4/3	16/9
Lines per frame	525	625	1125
Field Frequency (Hz)	60	50	60
Line Frequency (kHz)	15.75	15.625	33.75
Line Sweep time, μs	63.5	64	29.63
Video BW (MHz)	4.2	5.0	24.9
Sound	Mono/Stereo Output	Mono/Stereo Output	6-channel Dolby Digital Sound
Horizontal Retrace Time (μs)	10		3.7
Vertical Retrace, lines/field	21		45

4.8.1 NTSC TV Standard:

The baseband signal of NTSC standard needs a bandwidth of 4.5 MHz to transmit audio and video components of color TV. This corresponds to a 9.0 MHz bandwidth requirement for a DSB modulation scheme. Instead, a VSB+C technique -discussed in Section 4.5-- is used to fit the overall spectrum into a 6.0 MHz bandwidth allotted for each TV transmitter. The VHF and UHF station allocations and the key parameters of this standard are shown in Tables 4.4 and 4.5.

Table 4.4 VHF and UHF TV Station Frequency Allocations

VHF Ch: 2,3,4	54-60, 60-66, 66-72 MHz
VHF Ch: 5,6	76-82, 82-88 MHz
VHF Ch: 7-13	174-216 MH; 6 MHz Slots
UHF Ch: 14-83	470-890 MH; 6 MHz Slots

Note: Cable TV and Satellite TV channels have additional channels and the associated frequency assignments.

Table 4.5 NTSC North American TV Standards

Aspect ratio (width/height)	4/3
Lines per frame	525
Columns per line	525*(4/3)= 700
Total pixels per frame	367,500 pixels
Line rate	15,750 lines/sec.
Line duration	63.5 micro sec.
Horizontal retrace	10 micro sec.
Vertical retrace	1.27 msec.
Picture information per frame	485 lines

A picture element (pixel) is the smallest unit of an image that can be resolved. For instance, we have 367,500 pixels on a 525x700 NTSC standard. In other words, we can address individually that many points on the screen.

To produce an electrical signature of an image, the optical system in a camera generates a focused picture on a photo cathode of its tube, which produces an electrically charged image on the *target mosaic surface*. The electrical charges on this surface are then scanned by an electric gun, whose beam is controlled by voltages across horizontal and vertical deflection plates. This is called *imaging by* *orthicon*. Orthicon technique is now fairly old and obsolete, instead a less costly scheme called *imaging by vidicon* is the popular choice of the analog cameras in the market.

Yet, another important factor is the frame rate. Our visual perception system can treat a frame rate faster than 30 per second as a continuously moving picture. There are two methods of scanning moving pictures: (a) non-interlaced scanning with 30 frames per second, where the information on screen is cleared after every frame and (b) interlaced scanning, in which odd and even numbered fields (60 fields per second) are refreshed alternately as shown in Figure 4.28.

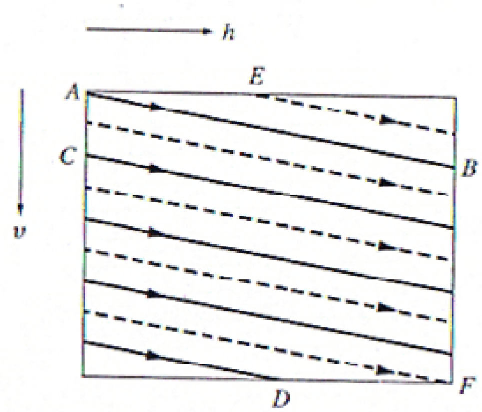


Figure 4.28 TV scanning raster in two fields.

Timing information in the form of horizontal and vertical blanking pulses is added to the video signal at the transmitter side to maintain proper line and field synchronization at the receiver. This composite video has a timing information. Timing diagram for the NTSC video standard is shown in Figure 4.29.

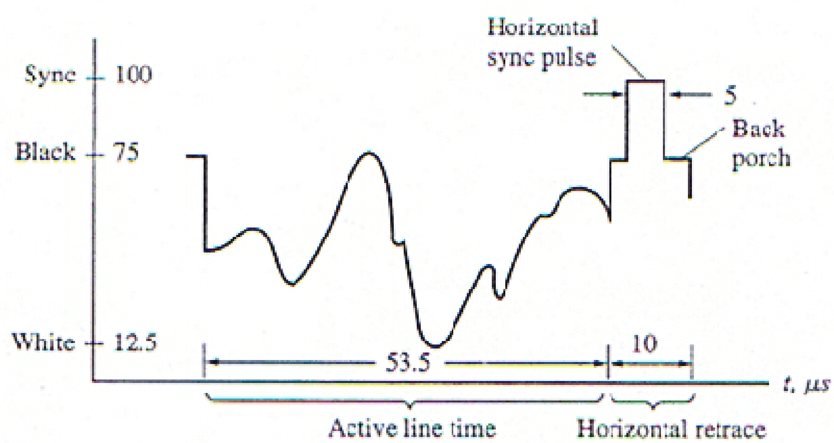


Figure 4.29 Video Timing Diagram for NTSC standard.

Blanking pul0ses of $10\mu s$ long are inserted during retrace intervals so that the receiver tube blanks out when the guns are retraced. Line synchronization pulse of $5\mu s$ are added on top of the blanking pulse to synchronize the horizontal sweep circuit of the receiver. A sequence of vertical synchronizing pulses is transmitted at the end of each field as the guns are vertically retraced back

Bandwidth of TV Signal: Let us now calculate the bandwidth of the video signal from the information content in the NTSC standard. During each frame or two fields 367,500 pixels are formed from 525 lines and 700 columns with aspect ration of 4/3. This corresponds to a transmission rate of 11,025,000 pixels per second. We will need a minimum bandwidth of 5.5 MHz by virtue of the Nyquist theorem. In moving pictures, adjacent pixels are highly correlated (redundant) and hence the actual bandwidth of the composite signal is less than that and 4.2 MHz is the norm used in commercial TV.

The audio signal is multiplexed with the video resulting in a composite signal with a total bandwidth of 4.5 MHz. This multiplexing is shown in Figure 4.30 overlapped with the linear vestigial side band (VSC) filter characteristic of the receiver.

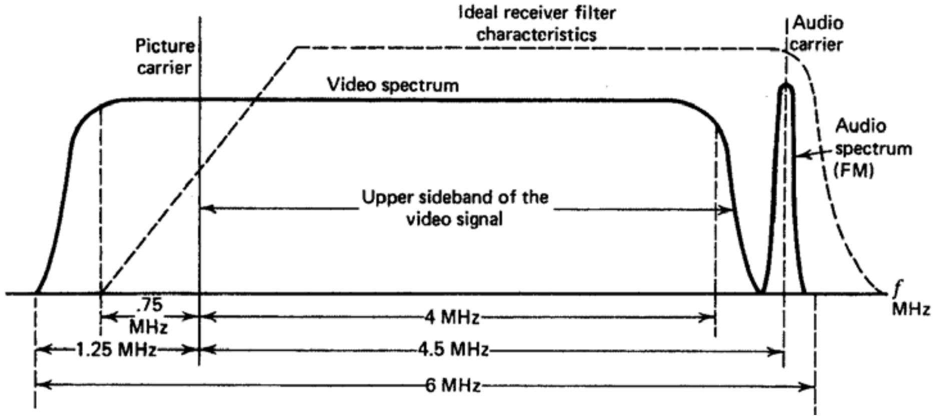
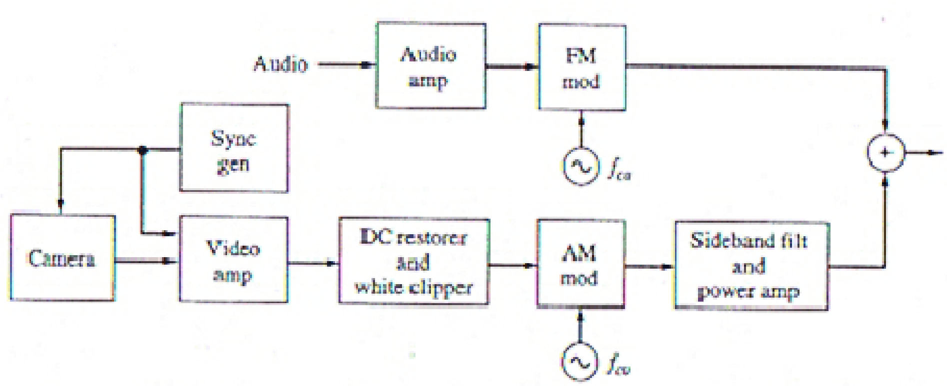


Figure 4.30. NTSC video spectrum including the stereo audio segment.

The upper sideband of the video is transmitted without attenuation up to 4.0 MHz with half-power frequency 4.2 MHz above the video carrier f_{vc} while the lower sideband has a 1.0 MHz bandwidth. Lower side band is left intact over the range 0-75 kHz and it is entirely attenuated at 1.25 MHz. The audio signal is FM modulated on a separate carrier $f_{ca} = f_{vc} + f_a$, with $f_a = 4.5 MHz$ and frequency deviation of $f_\Delta = 25\ kHz$ as it will be discussed later resulting in signal with 80 kHz bandwidth. Thus, the total bandwidth of the modulated composite signal amounts to 5.75 MHz leaving a 250 kHz guard band between adjacent TV channels. The transmitter and receiver block diagrams for a monochrome (B/W) TV are shown in Figure 4.31 and Figure 4.32, respectively..

B/W TV Transmitter



B/W TV Receiver

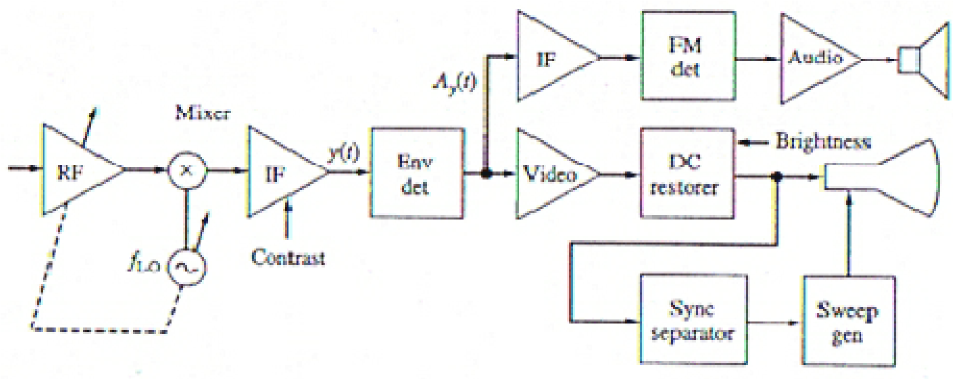


Figure 4.31-4.32. Transmitter and receiver block diagrams of the NTSC mono TV system.

The signal at the input to the envelope detector in the receiver is given by:

$$y(t) = A_{cv}[1+\mu].Cosw_{vc}t - A_{vc}.\mu.x_q(t)Sinw_{vc}t + A_{ca}Cos[(w_{vc}+w_a)t+\phi(t)] \qquad (4.54)$$

The video is modulated in AM with $\mu = 0.875$, where $x(t)$ is the video signal, $\phi(t)$ is the FM audio with constraints: $|\mu.x_q(t)| << 1$ and $A_{ca} << A_{cv}$ which results in an envelope signal:

$$A_y(t) = A_{cv}[1+\mu.x(t)] + A_{ca}.Cos[w_a t + \phi(t)] \qquad (4.55)$$

Hence, the receiver structure in Figure 4.32 will work successfully.

4.8.2 Color Processing:

Color imagery can be treated either in the Red, Green, and Blue (RGB) or the luminance-chrominance space. The ubiquitous color chart in Figure 4.33 is a very familiar figure to people working with desktop screens. Each point on this color chart can be expressed in terms of R, G, and B values in the range 0 to 15 for a 4-bit per color resolution or 0 to 255 per each color of 8-bit per color or 24-bit total resolution. The set of two pictures in Figure 4.34 depicts the effects of high resolution imagery (8-bits per color, 24-bit per pixel) versus low resolution imagery (4-bits per color, 12-bit per pixel).

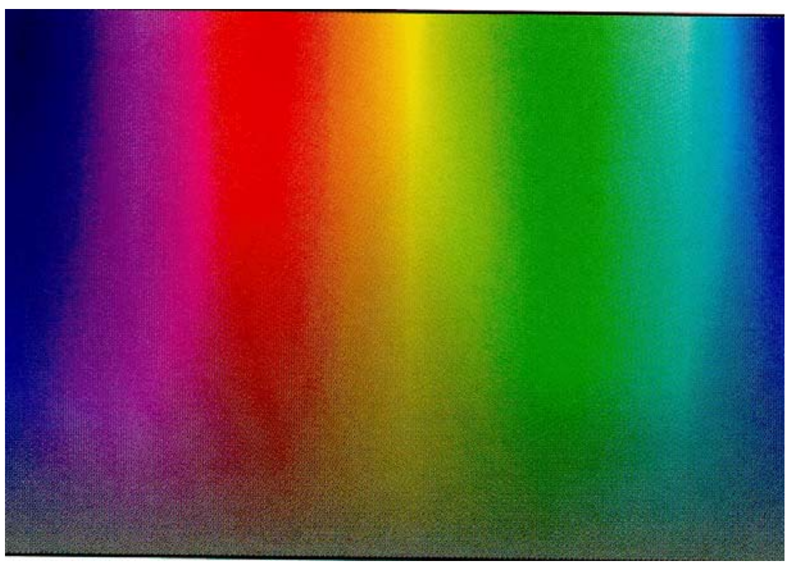


Figure 4.33. Color chart.

Figure 4.34. 24-bits per pixel and 12-bits per pixel imagery.

But, {R, G, B} color representation is not used in transmitting video information. Instead, more efficient color representations are used. The luminance represents the intensity (brightness) in an image pixel as the case of monotonic (B/W) TV. The luminance based on response measurements from eye is given by:

$$m_L(t) = 0.30m_r(t) + 0.59m_g(t) + 0.11m_b(t) \tag{4.56}$$

where

$m_r(t)$ ≡ red camera signal
$m_g(t)$ ≡ green camera signal
$m_b(t)$ ≡ blue camera signal

and

$$m_I(t) = 0.60m_r(t) - 0.28m_g(t) - 0.32m_b(t) \tag{4.57}$$

$$m_Q(t) = 0.21m_r(t) - 0.52m_g(t) + 0.31m_b(t) \tag{4.58}$$

The remaining two signals, called chrominance signals, contain additional information that can be used to reconstruct $\{m_r(t), m_g(t), m_b(t)\}$ from which the original color can be synthesized.

The luminance signal $m_L(t)$ is allotted 4.2 MHz base-band bandwidth and transmitted as VSB as in the B/W TV. The chrominance signals are quadrature multiplexed on a color sub-carrier at $f_C + 3{,}579{,}545\ Hz$, which falls exactly between the 227th and 228th harmonic of the line frequency. The base-band color signal has the form:

$$x_b(t) = m_L(t) + m_Q(t)Sinw_Ct + m_I(t)Cosw_{cc}t + m_I(t)Sinw_{cc}t \tag{4.59}$$

where w_{cc} is the color sub-carrier frequency. Horizontal and vertical synchronizing pulses are added to $x_b(t)$ at the transmitter before piping it out to the channel. Additionally, an eight-cycle piece of the color sub-carrier known as the “color burst” is put on the “back porch” of the blanking pulses for color sub-carrier synchronization at the receiver.

At the receiver, the video signal $x_b(t)$ is recovered by an enveloped demodulator. The luminance signal is handled directly except amplification since it is a base-band signal. A synchronous color carrier is generated from the color burst and the chrominance components are synchronously demodulated. The recovered luminance and chrominance information is linearly converted into R,G,B components by the following matrix:

$$m_r(t) = x_L(t) - 0.96x_I(t) + 0.62x_Q(t) \tag{4.60a}$$

$$m_g(t) = x_L(t) - 0.28x_I(t) - 0.64x_Q(t) \tag{4.60b}$$

$$m_b(t) = x_L(t) - 1.10x_I(t) + 1.70x_Q(t) \tag{4.60c}$$

depending on the picture tube used in the receiver, these primary color signals are combined in different ways to produce a close replica of the original analog picture.

When a color signal is applied to a monochrome receiver, the viewer does not see the sinusoidal variations produced by the color subcarrier and its sidebands. If a B/W signal is applied to a color TV receiver, all three color signals will are assumed to be equal and the reconstructed signal become a B/W picture.