Detection of SSB Signals with a Carrier (SSB+C)

We now consider SSB signals with an additional carrier (SSB+C). Such a signal can be expressed as

$$\varphi_{\text{SSB+C}} = A \cos \omega_c t + [m(t) \cos \omega_c t + m_h(t) \sin \omega_c t]$$

and $m(t)$ can be recovered by synchronous detection [multiplying $\varphi_{\text{SSB+C}}$ by $\cos \omega_c t$] if the carrier component $A \cos \omega_c t$ can be extracted (by narrowband filtering of) $\varphi_{\text{SSB+C}}$. Alternatively, if the carrier amplitude A is large enough, $m(t)$ can also be (approximately) recovered from $\varphi_{\text{SSB+C}}$ by envelope or rectifier detection. This can be shown by rewriting $\varphi_{\text{SSB+C}}$ as

$$\begin{aligned} \varphi_{\text{SSB+C}} &= [A + m(t)] \cos \omega_c t + m_h(t) \sin \omega_c t \\ &= E(t) \cos (\omega_c t + \theta) \end{aligned} \tag{4.21}$$

where $E(t)$, the envelope of $\varphi_{\text{SSB+C}}$, is given by [see Eq. (3.41a)]

$$\begin{aligned} E(t) &= \{[A + m(t)]^2 + m_h^2(t)\}^{1/2} \\ &= A\left[1 + \frac{2m(t)}{A} + \frac{m^2(t)}{A^2} + \frac{m_h^2(t)}{A^2}\right]^{1/2} \end{aligned}$$

If $A \gg |m(t)|$, then in general* $A \gg |m_h(t)|$, and the terms $m^2(t)/A^2$ and $m_h^2(t)/A^2$ can be ignored. Thus,

$$E(t) \simeq A\left[1 + \frac{2m(t)}{A}\right]^{1/2}$$

Using Taylor series expansion and discarding higher order terms [because $m(t)/A \ll 1$], we get

$$\begin{aligned} E(t) &\simeq A\left[1 + \frac{m(t)}{A}\right] \\ &= A + m(t) \end{aligned}$$

It is evident that for a large carrier, the SSB + C can be demodulated by an envelope detector.

In AM, envelope detection requires the condition $A \geq |m(t)|$, whereas for SSB+C, the condition is $A \gg |m(t)|$. Hence, in SSB case, the required carrier amplitude is much larger than that in AM, and, consequently, the efficiency of SSB+C is pathetically low.

Quadrature Amplitude Modulation (QAM)

Because SSB-SC signals are difficult to generate accurately, quadrature amplitude modulation (QAM) offers an attractive alternative to SSB-SC. QAM can be exactly generated without requiring sharp-cutoff bandpass filters. QAM operates by transmitting two DSB signals using carriers of the same frequency but in phase quadrature, as shown in Fig. 4.19. This scheme is known as **quadrature amplitude modulation (QAM)** or **quadrature multiplexing**.

As shown Figure 4.19, the boxes labeled $-\pi/2$ are phase shifters that delay the phase of an input sinusoid by $-\pi/2$ rad. If the two baseband message signals for transmission are $m_1(t)$ and $m_2(t)$, the corresponding QAM signal $\varphi_{\text{QAM}}(t)$, the sum of the two DSB-modulated signals, is

$$\varphi_{\text{QAM}}(t) = m_1(t) \cos \omega_c t + m_2(t) \sin \omega_c t$$

* This may not be true for all t, but it is true for most t.

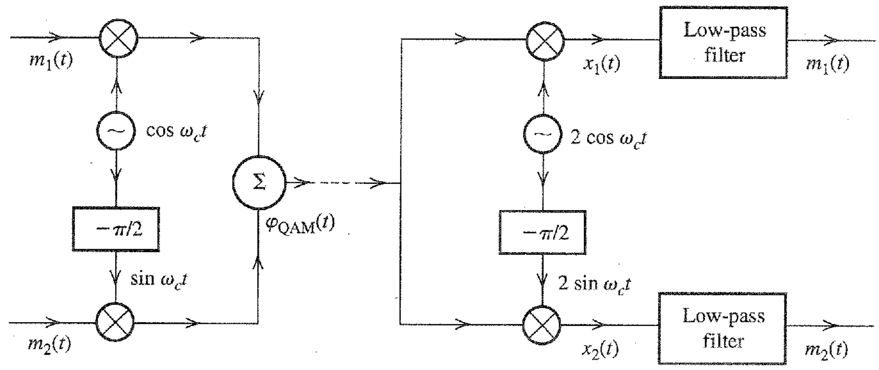

Figure 4.19 Quadrature amplitude multiplexing.

Both modulated signals occupy the same band. Yet two baseband signals can be separated at the receiver by synchronous detection if two local carriers are used in phase quadrature, as shown in Fig. 4.19. This can be shown by considering the multiplier output $x_1(t)$ of the upper arm of the receiver (Fig. 4.19):

$$\begin{aligned} x_1(t) &= 2\varphi_{\text{QAM}}(t) \cos \omega_c t = 2[m_1(t) \cos \omega_c t + m_2(t) \sin \omega_c t] \cos \omega_c t \\ &= m_1(t) + m_1(t) \cos 2\omega_c t + m_2(t) \sin 2\omega_c t \end{aligned} \tag{4.22a}$$

The last two terms are bandpass signals centered around $2\omega_c$. In fact, they actually form a QAM signal with $2\omega_c$ as the carrier frequency. They are suppressed by the low-pass filter, yielding the desired demodulation output $m_1(t)$. Similarly, the output of the lower receiver branch can be shown to be $m_2(t)$.

$$\begin{aligned} x_2(t) &= 2\varphi_{\text{QAM}}(t) \sin \omega_c t = 2[m_1(t) \cos \omega_c t + m_2(t) \sin \omega_c t] \sin \omega_c t \\ &= m_2(t) - m_2(t) \cos 2\omega_c t + m_1(t) \sin 2\omega_c t \end{aligned} \tag{4.22b}$$

Thus, two baseband signals, each of bandwidth B Hz, can be transmitted simultaneously over a bandwidth $2B$ by using DSB transmission and quadrature multiplexing. The upper channel is also known as the **in-phase** (I) channel and the lower channel is the **quadrature** (Q) channel. Both signals $m_1(t)$ and $m_2(t)$ can be separately demodulated.

Note, however, that QAM demodulation must be totally synchronous. An error in the phase or the frequency of the carrier at the demodulator in QAM will result in loss and interference between the two channels. To show this, let the carrier at the demodulator be $2 \cos(\omega_c t + \theta)$. In this case,

$$\begin{aligned} x_1(t) &= 2[m_1(t) \cos \omega_c t + m_2(t) \sin \omega_c t] \cos(\omega_c t + \theta) \\ &= m_1(t) \cos\theta - m_2(t) \sin\theta + m_1(t) \cos(2\omega_c t + \theta) + m_2(t) \sin(2\omega_c t + \theta) \end{aligned}$$

The low-pass filter suppresses the two signals modulated by carrier of angular frequency $2\omega_c$, resulting in the first demodulator output

$$m_1(t) \cos\theta - m_2(t) \sin\theta$$

Thus, in addition to the desired signal $m_1(t)$, we also receive signal $m_2(t)$ in the upper receiver branch. A similar phenomenon can be shown for the lower branch. This so-called **cochannel interference** is undesirable. Similar difficulties arise when the local frequency is in error (see

Prob. 4.4-1). In addition, unequal attenuation of the USB and the LSB during transmission leads to cross talk or cochannel interference.

Quadrature multiplexing is used in analog color television to multiplex the so-called chrominance signals, which carry the information about colors. There, the synchronization is achieved by periodic insertion of a short burst of carrier signal (called **color burst** in the transmitted signal). Digital satellite television transmission also applies QAM.

In terms of bandwidth requirement, SSB is similar to QAM but less exacting in terms of the carrier frequency and phase or the requirement of a distortionless transmission medium. However, SSB is difficult to generate if the baseband signal $m(t)$ has significant spectral content near the dc.

4.5 AMPLITUDE MODULATIONS: VESTIGIAL SIDEBAND (VSB)

As discussed earlier, it is rather difficult to generate exact SSB signals. They generally require that the message signal $m(t)$ have a null around dc. A phase shifter, required in the phase shift method, is unrealizable, or only approximately realizable. The generation of DSB signals is much simpler, but it requires twice the signal bandwidth. **Vestigial sideband (VSB)** modulation, also called the asymmetric sideband system, is a compromise between DSB and SSB. It inherits the advantages of DSB and SSB but avoids their disadvantages at a small cost. VSB signals are relatively easy to generate, and, at the same time, their bandwidth is only a little (typically 25%) greater than that of SSB signals.

In VSB, instead of rejecting one sideband completely (as in SSB), a gradual cutoff of one sideband as shown in Fig. 4.20d, is accepted. The baseband signal can be recovered exactly by a synchronous detector in conjunction with an appropriate equalizer filter $H_o(f)$ at the receiver output (Fig. 4.21). If a large carrier is transmitted along with the VSB signal, the baseband signal can be recovered by an envelope (or a rectifier) detector.

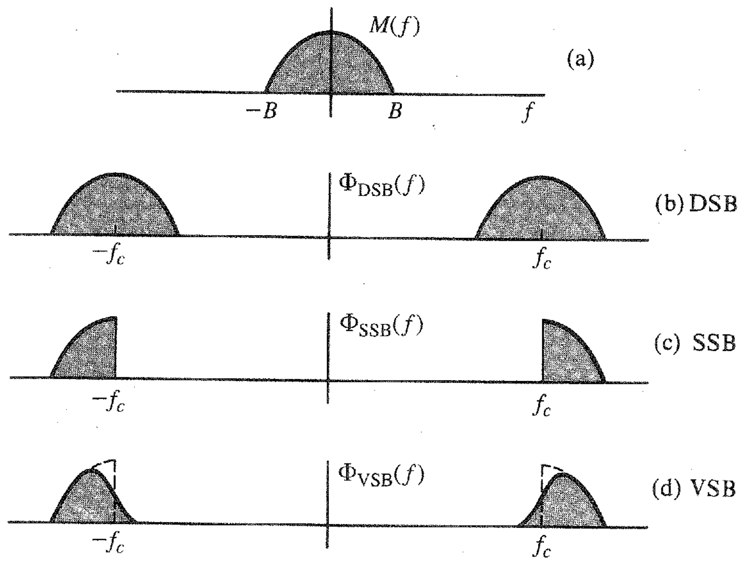

Figure 4.20 Spectra of the modulating signal and corresponding DSB, SSB, and VSB signals.

If the vestigial shaping filter that produces VSB from DSB is $H_i(f)$ (Fig. 4.21), then the resulting VSB signal spectrum is

$$\Phi_{\text{VSB}}(f) = [M(f+f_c) + M(f-f_c)]H_i(f) \tag{4.23}$$

This VSB shaping filter $H_i(f)$ allows the transmission of one sideband but suppresses the other sideband, not completely, but gradually. This makes it easy to realize such a filter, but the transmission bandwidth is now somewhat higher than that of the SSB (where the other sideband is suppressed completely). The bandwidth of the VSB signal is typically 25 to 33% higher than that of the SSB signals.

We require that $m(t)$ be recoverable from $\varphi_{\text{VSB}}(t)$ by using synchronous demodulation at the receiver. This is done by multiplying the incoming VSB signal $\varphi_{\text{VSB}}(t)$ by $2\cos\omega_c t$. The product $e(t)$ is given by

$$e(t) = 2\varphi_{\text{VSB}}(t)\cos\omega_c t \Longleftrightarrow [\Phi_{\text{VSB}}(f+f_c) + \Phi_{\text{VSB}}(f-f_c)]$$

The signal $e(t)$ is further passed through the low-pass equalizer filter of the transfer function $H_o(f)$. The output of the equalizer filter is required to be $m(t)$. Hence, the output signal spectrum is given by

$$M(f) = [\Phi_{\text{VSB}}(f+f_c) + \Phi_{\text{VSB}}(f-f_c)]H_o(f)$$

Substituting Eq. (4.23) into this equation and eliminating the spectra at $\pm 4f_c$ [suppressed by a low-pass filter $H_o(f)$], we obtain

$$M(f) = M(f)[H_i(f+f_c) + H_i(f-f_c)]H_o(f) \tag{4.24}$$

Hence

$$H_o(f) = \frac{1}{H_i(f+f_c) + H_i(f-f_c)} \qquad |f| \le B \tag{4.25}$$

Note that because $H_i(f)$ is a bandpass filter, the terms $H_i(f \pm f_c)$ contain low-pass components.

Complementary VSB Filter and Envelope Detection of VSB + C Signals

As a special case of a filter at the VSB modulator, we can choose $H_i(f)$ such that

$$H_i(f+f_c) + H_i(f-f_c) = 1 \qquad |f| \le B \tag{4.26}$$

The output filter is just a simple low-pass filter with transfer function:

$$H_o(f) = 1 \qquad |f| \le B$$

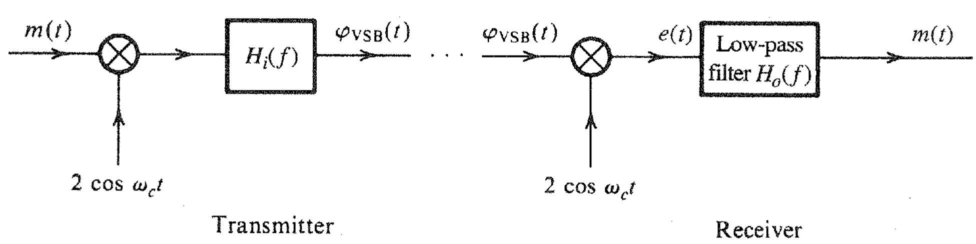

Figure 4.21 VSB modulator and demodulator.

The resulting VSB signal plus carrier (VSB + C) can be envelope-detected. This demodulation method may be proved by using exactly the same argument used in proving the case for SSB + C signals. In particular, because of Eq. (4.26), we can define a new low-pass filter

$$F(f) = j[1 - 2H_i(f - f_c)] = -j[1 - 2H_i(f + f_c)] \qquad |f| \le B$$

Defining a new (complex) low-pass signal as

$$m_v(t) \Longleftrightarrow M_v(f) = F(f)M(f)$$

we can rewrite the VSB signal as

$$\Phi_{\mathrm{VSB}}(f) = \frac{M(f - f_c) + M(f + f_c)}{2} + \frac{M_v(f - f_c) - M_v(f + f_c)}{2j} \tag{4.27a}$$

$$\Longleftrightarrow$$

$$\varphi_{\mathrm{VSB}}(t) = m(t)\cos 2\pi f_c t + m_v(t)\sin 2\pi f_c t \tag{4.27b}$$

Clearly, both the SSB and the VSB modulated signals have the same form, with $m_h(t)$ in SSB replaced by a low-pass signal $m_v(t)$ in VSB. Applying the same analysis from the SSB+C envelope detection, a large carrier addition to $\varphi_{\mathrm{VSB}}(t)$ would allow the envelope detection of VSB + C.

We have shown that SSB+C requires a much larger carrier than DSB+C (AM) for envelope detection. Because VSB+C is an in-between case, the added carrier required in VSB is larger than that in AM, but smaller than that in SSB + C.

Example 4.7 The carrier frequency of a certain VSB signal is $f_c = 20$ kHz, and the baseband signal bandwidth is 6 kHz. The VSB shaping filter $H_i(f)$ at the input, which cuts off the lower sideband gradually over 2 kHz, is shown in Fig. 4.22a. Find the output filter $H_o(f)$ required for distortionless reception.

Figure 4.22b shows the low-pass segments of $H_i(f + f_c) + H_i(f - f_c)$. We are interested in this spectrum only over the baseband (the remaining undesired portion is suppressed by the output filter). This spectrum, which is 0.5 over the band of 0 to 2 kHz, is 1 from 2 to 6 kHz, as shown in Fig. 4.22b. Figure 4.22c shows the desired output filter $H_o(f)$, which is the reciprocal of the spectrum in Fig. 4.22b [see Eq. (4.25)].

Use of VSB in Broadcast Television

VSB is a clever compromise between SSB and DSB, which makes it very attractive for television broadcast systems. The baseband video signal of television occupies an enormous bandwidth of 4.5 MHz, and a DSB signal needs a bandwidth of 9 MHz. It would seem desirable to use SSB to conserve bandwidth. Unfortunately, doing this creates several problems. First, the baseband video signal has sizable power in the low-frequency region, and consequently it is difficult to suppress one sideband completely. Second, for a broadcast receiver, an envelope detector is preferred over a synchronous one to reduce the receiver cost. We saw earlier that SSB+C has a very low power efficiency. Moreover, using SSB will increase the receiver cost.

The spectral shaping of television VSBs signals can be illustrated by Fig. 4.23. The vestigial spectrum is controlled by two filters: the transmitter RF filter $H_T(f)$ and the receiver RF filter

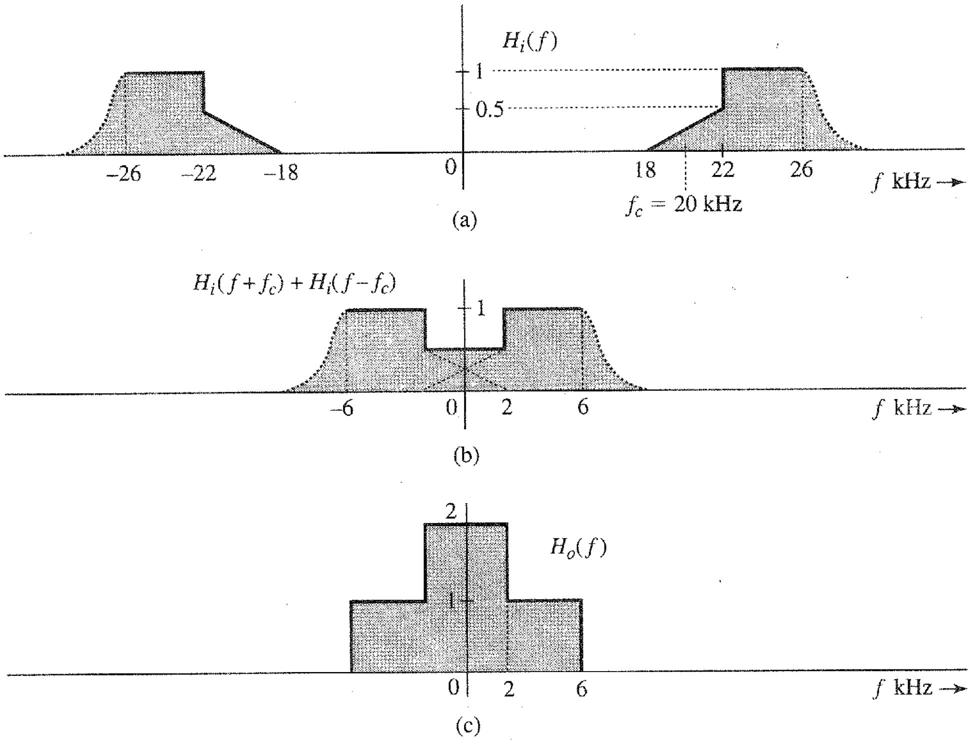

Figure 4.22 VSB modulator and receiver filters.

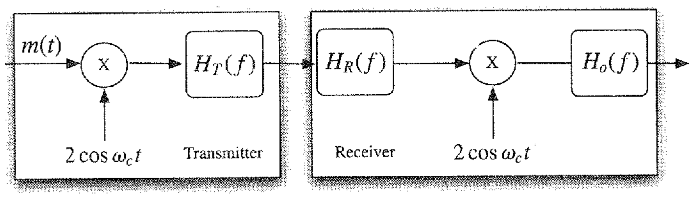

Figure 4.23 Transmitter filter $H_T(f)$, receiver front-end filter $H_R(f)$, and the receiver output low-pass filter $H_o(f)$ in VSB television systems.

$H_R(f)$. Jointly we have

$$H_i(f) = H_T(f)H_R(f)$$

Hence, the design of the receiver output filter $H_o(f)$ follows Eq. (4.25).

The DSB spectrum of a television signal is shown in Fig. 4.24a. The vestigial shaping filter $H_i(f)$ cuts off the lower sideband spectrum gradually, starting at 0.75 MHz to 1.25 MHz below the carrier frequency f_c, as shown in Fig. 4.24b. The receiver output filter $H_o(f)$ is designed according to Eq. (4.25). The resulting VSB spectrum bandwidth is 6 MHz. Compare this with the DSB bandwidth of 9 MHz and the SSB bandwidth of 4.5 MHz.

4.6 LOCAL CARRIER SYNCHRONIZATION

In a suppressed carrier, amplitude-modulated system (DSB-SC, SSB-SC, and VSB-SC), the coherent receiver must generate a local carrier that is synchronous with the incoming carrier (frequency and phase). As discussed earlier, any discrepancy in the frequency or phase of the local carrier gives rise to distortion in the detector output.