Double-Side Band Suppressed Carrier (DSBSC) Modulation

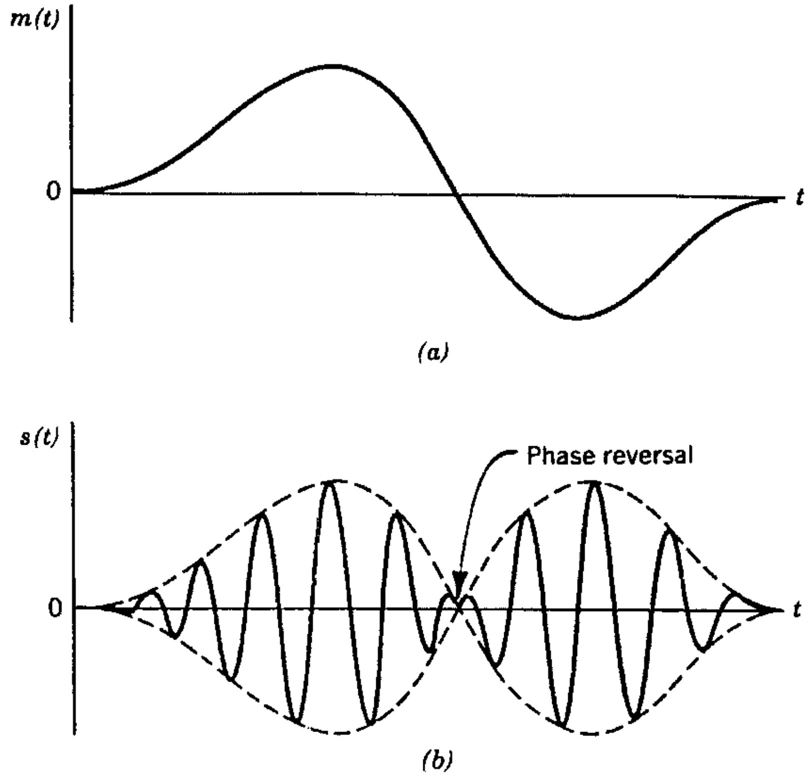
With ordinary AM, we have seen that a significant amount of power in the transmitted signal is wasted in the carrier. The benefit of this is a very simple cheap receiver.

DSBSC modulation is a modulation scheme that is more power efficient than AM, yet requires a more complex receiver.

The concept of DSBSC is extremely simple. To modulate a message $m(t)$, we simply multiply it by a sinusoid at freq f_c. The modulated wave $s(t)$ is given as

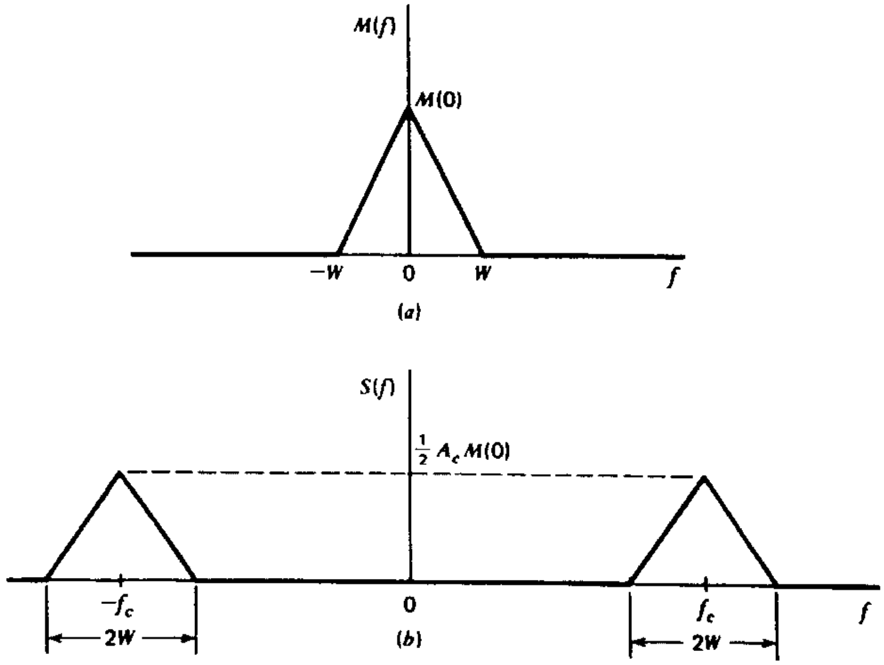
$$s(t) = m(t)\, c(t)$$
$$= A_c\, m(t) \cos 2\pi f_c t \qquad \longrightarrow (1)$$

Where $c(t) = A_c \cos 2\pi f_c t$ is the carrier.



(a) *Message signal.* (b) *DSBSC-modulated wave* s(t).

Fig. 8a



(a) *Spectrum of message signal.* (b) *Spectrum of DSBSC modulated wave.*

Fig. 8b.

From our previous discussion on multiplication by sinusoids, we know the modulated spectrum $S(f)$ is a translated version of $M(f)$. The resulting time and freq. domain rep'ns of the DSBSC wave are shown in Fig. 8.

We see that, unlike AM, the carrier component is not present, so the power efficiency for DSB-SC is improved over AM. However, to recover $m(t)$, we must multiply the received signal $s(t)$ again by $c(t) = \cos 2\pi f_c t$, i.e.,

$$s'(t) = s(t) \cdot \cos 2\pi f_c t$$

$$= A_c\, m(t) \cos^2 (2\pi f_c t)$$

$$= \frac{A_c m(t)}{2} \cdot \left[1 + \cos (2\pi \cdot 2 f_c \cdot t)\right] \rightarrow (2)$$

The original signal can be recovered by using the low pass filter which reject the components at $2f_c$.

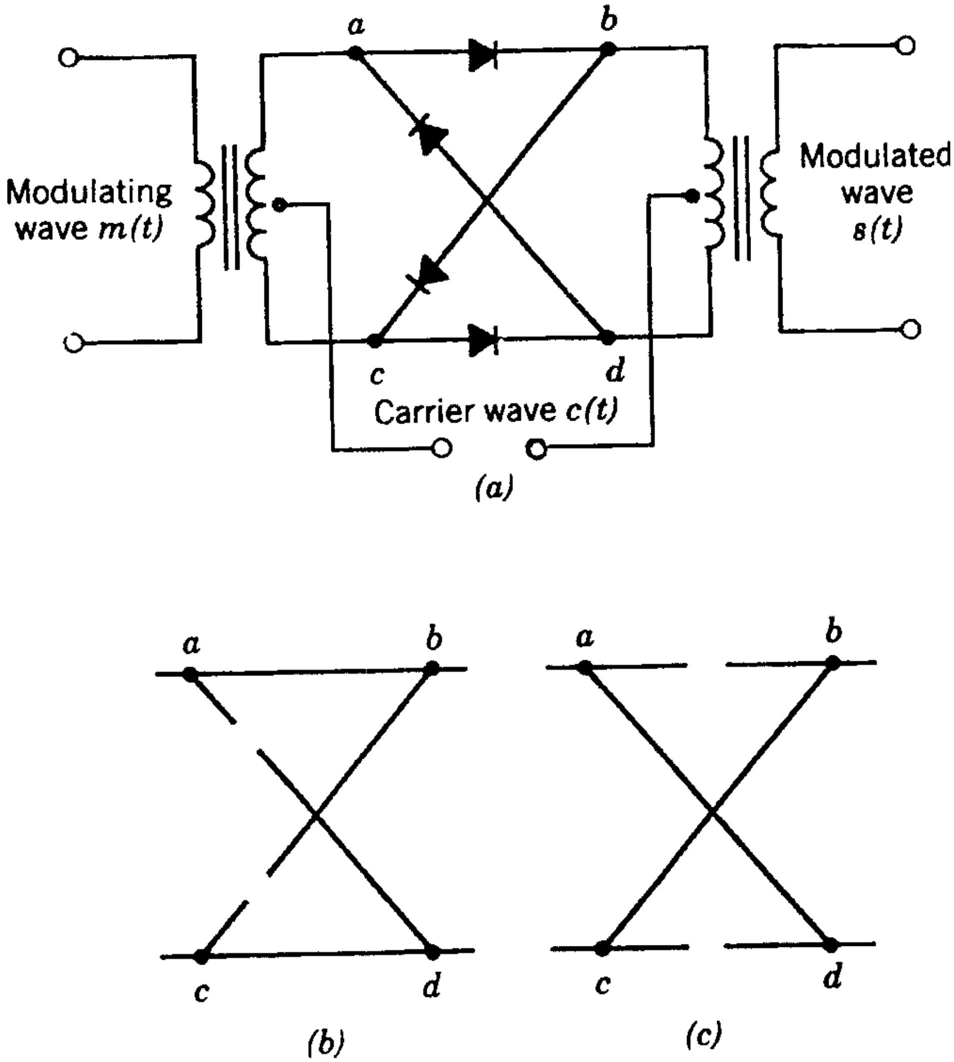


Fig 9

Ring modulator. (a) *Circuit diagram.* (b) *The condition when the outer diodes are switched on and the inner diodes are switched off.* (c) *The condition when the outer diodes are switched off and the inner diodes are switched on.*

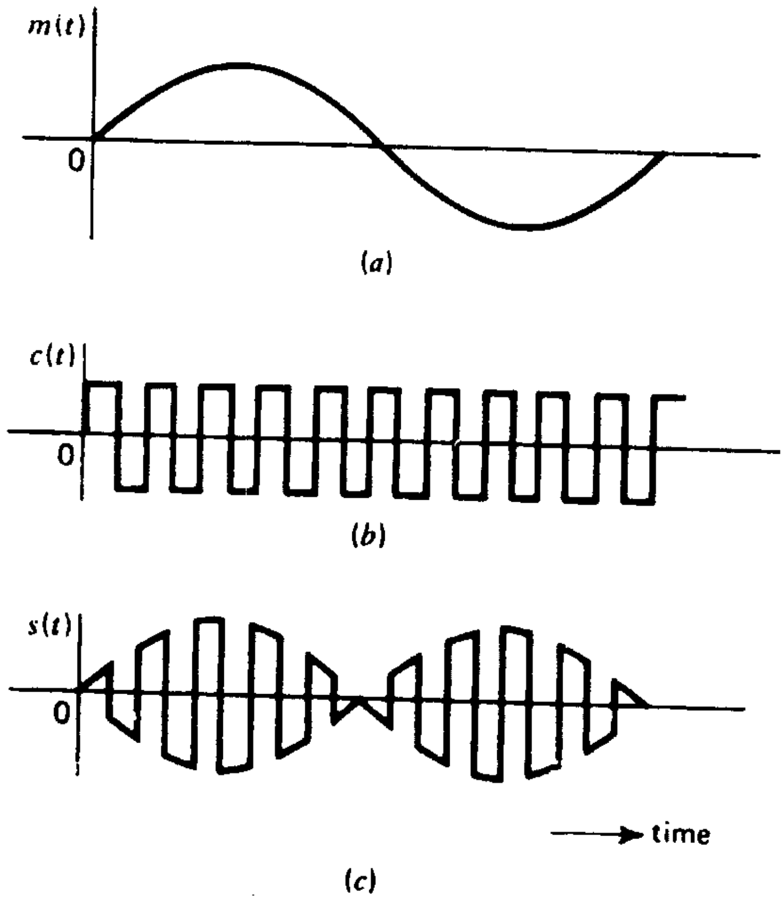


Fig. 10

Waveforms illustrating the operation of the ring modulator for a sinusoidal modulating wave. (a) *Modulating wave.* (b) *Square-wave carrier.* (c) *Modulated wave.*

Generation of DSBSC signals.

This is commonly done using a double-balanced mixer, or a ring modulation shown in Fig. 9a.

The carrier $c(t)$ is large compared to $m(t)$. Its job is to switch the diode pairs on and off. On the positive ½-cycle of $c(t)$, $V_c > V_d$ and $V_a > V_b$, and therefore, the crossed diodes are off, and the connection between the in and output transformer is shown in Fig. 9b. Thus, $m(t)$ arrives at the output transformer with +ve polarity. On the -ve ½-cycle of $c(t)$, $V_c < V_d$ and $V_a < V_b$, therefore, the crossed diodes are on (the "thru" diodes are now off). The effective connection between the transformers is shown in Fig. 9c. Now, $m(t)$ arrives at the output with -ve polarity. The result of this circuit is that the signal $m(t)$ gets "chopped" into + and - segments as shown in Fig. 10.

Because $c(t)$ switches the diodes abruptly, the modulated signal $s(t)$ is effectively $m(t)$ multiplied by a square wave at f_0. thus, $s(t)$ produced from the ring modulator is given by:

$$s(t) = m(t)\, c(t)$$

where $c(t)$ is now a square wave with duty cycle $\frac{1}{2}$ with p-p amplitude 1 at the carrier frequency. Thus, $c(t)$ can be expresed as a Fourier series:

$$c(t) = \sum_{n=\infty}^{\infty} c_n \cdot e^{j2\pi n t f_c} \longrightarrow (3)$$

since $c_n = c_{-n}$ for the symmetric square wave,

$$c(t) = c_0 + \sum_{n=1}^{\infty} c_n \cdot \left[e^{j2\pi n f_c t} + e^{-j2\pi n f_c t} \right]$$

$$= c_0 + 2\sum_{n=1}^{\infty} c_n \cdot \cos(2\pi n f_c t) \longrightarrow (4)$$

For the square wave shown in Fig. 10b, the average or DC component is zero, i.e. $c_0 = 0$, Also, even harmonics are zero. Therefore (4) can be rewritten as,

$$c(t) = 2\sum_{m=1}^{\infty} c_{2m-1} \cdot \cos\left[2\pi (2m-1) f_c t\right] \longrightarrow (5)$$

and $s(t)$ is therefore:

$$s(t) = \sum_{n=1}^{\infty} c_{2m-1} \cos\left[2\pi f_c t\,(2m-1)\right] \times m(t) \rightarrow (6)$$

The multiplication of $m(t)$ by $\cos[2\pi f_c t(2m-1)]$ results in the shift of the spectral components to $\pm(2m-1)f_c$. Therefore, we get the spectral components centered at $\pm f_c$, $\pm 3f_c$, $\pm 5f_c$ and so on.

By using a band pass filter with the centre frequency f_c, we can obtain ~~et~~ the desired DSBSC signal.

If the signal is bandlimited to w Hz, the bandwidth of the bandpass filter need not be larger than w Hz. Since $w < f_c$, this bandpass filter rejects the components at $\pm 3f_c$, $\pm 5f_c$ and so on

Coherent Detection of DSBSC Waves.

We have seen from the discussion pg.2, that to recover a message signal $m(t)$ from a DSBSC wave, we multiply $m(t)$ again by $c(t) = \cos 2\pi f_0 t$.
i.e. let $s(t)$ be

$$s(t) = m(t) \cos 2\pi f_0 t \quad \text{(DSBSC wave)} \tag{7}$$

To recover $m(t)$, we multiply $s(t)$ again by $c(t) = \cos 2\pi f_0 t$. But generally, we can not reconstruct $c(t) = \cos (2\pi f_0 t)$ at the receiver (for this, we need a costas loop) Note that for perfect demodulation, the demodulating $c(t)$ must be perfectly matched to $\cos 2\pi f_0 t$ both in freq. and phase.

Let us examine what happens when we demodulate $s(t)$ above with $\cos(2\pi f_0 t + \phi)$ — i.e., $c(t)$ now contains a phase error:

$$\hat{s}(t) = s(t)\cos(2\pi f_0 t + \phi) \quad (8)$$
$$= m(t)\cos(2\pi f_0 t)\cos(2\pi f_0 t + \phi)$$
$$= \tfrac{1}{2} m(t)\cos\phi + \underbrace{\tfrac{1}{2}\cos(4\pi f_0 t + \phi)\, m(t)}$$

unwanted term may be removed by filtering.

The output of the multiplier is thus

$$\hat{s}_0(t) = \tfrac{1}{2} m(t)\cos\phi. \quad (9)$$

When the phase error ϕ is small, $\hat{s}_0(t)$ is a scaled version of $m(t)$, since $\cos\phi \sim 1$.

But if $\phi \to 90^\circ$, severe attenuation in the recovered signal can result.