DEPARTMENT OF ECE

ANALOG COMMMUNICATION
UNIT-I
DOUBLE SIDE BAND - SUPRESSED CARRIER

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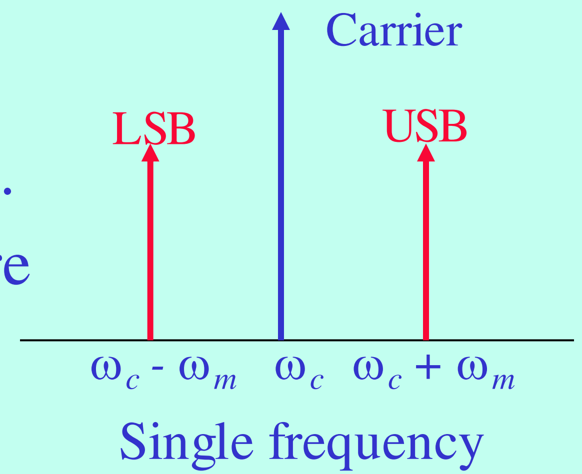
Contents

Double Side Band Suppressed Carrier



From AM spectrum:

- Carries signal ω_c carries no information ω_m.
- Carries signal consumes a lot of power more than 50%

Question: *Why transmit carrier at all?*

Ans:

Question: *Can one suppress the carrier?*

Ans.: *Yes, just transmit two side bands (i.e* ***DSB-SC****)*

But what is the penalty?

System complexity at the receiver

DSB-SC - *Theory*

General expression: $c(t) = [k_1 m(t) + C] \cos(\omega_c t + \phi_c)$

Let $k_1 = 1$, $C = 0$ and $\phi_c = 0$, the modulated carrier signal, therefore:

$$c(t) = m(t) \cos \omega_c t$$

Information signal $m(t) = E_m \cos \omega_m t$

Thus

$$c(t) = E_m \cos \omega_m t \cos \omega_c t$$

$$= \frac{ME_c}{2} \cos(\omega_c + \omega_m)t + \frac{ME_c}{2} \cos(\omega_c - \omega_m)t$$

upper side band

lower side band

DSB-SC - *Waveforms*

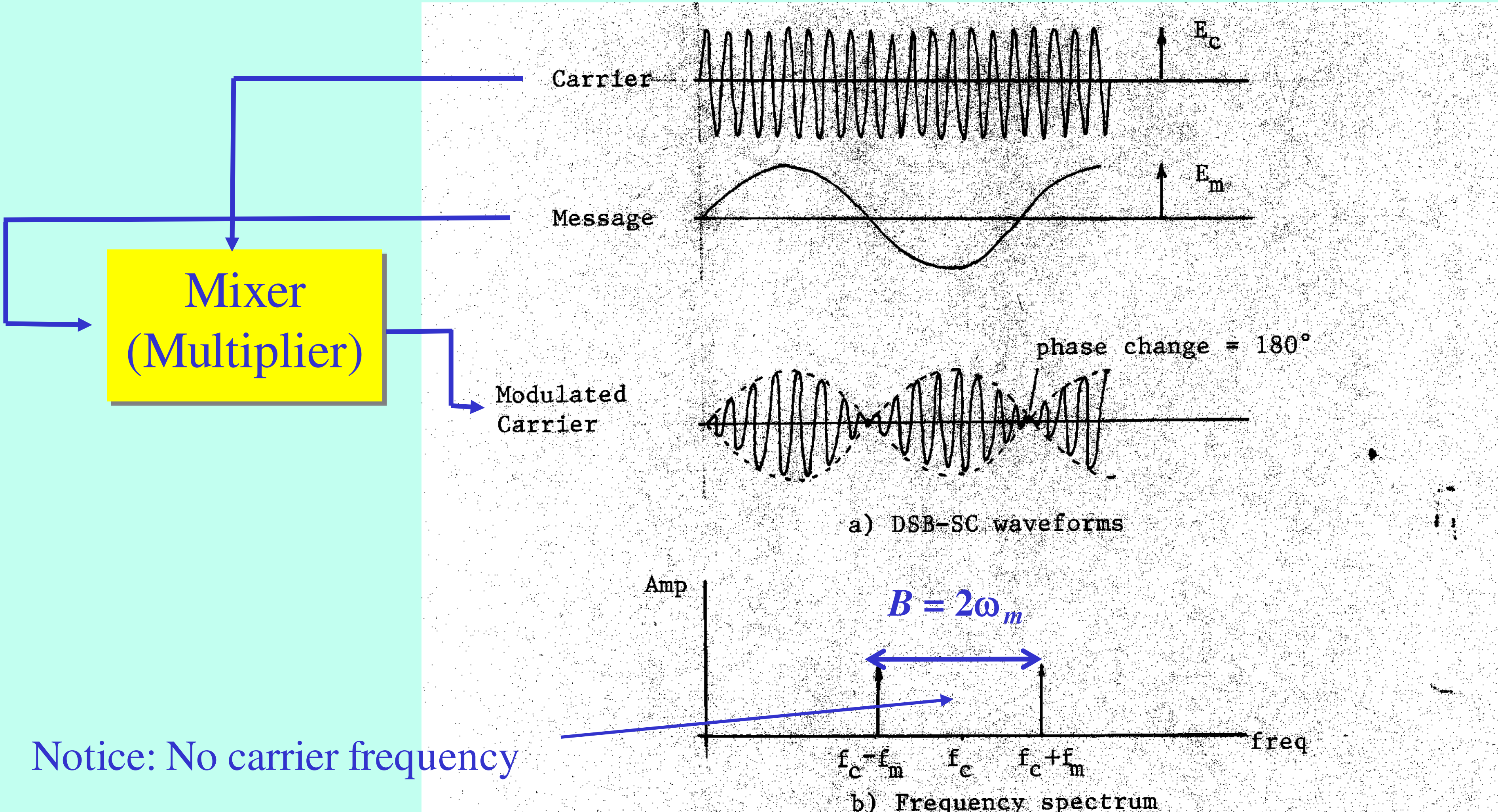
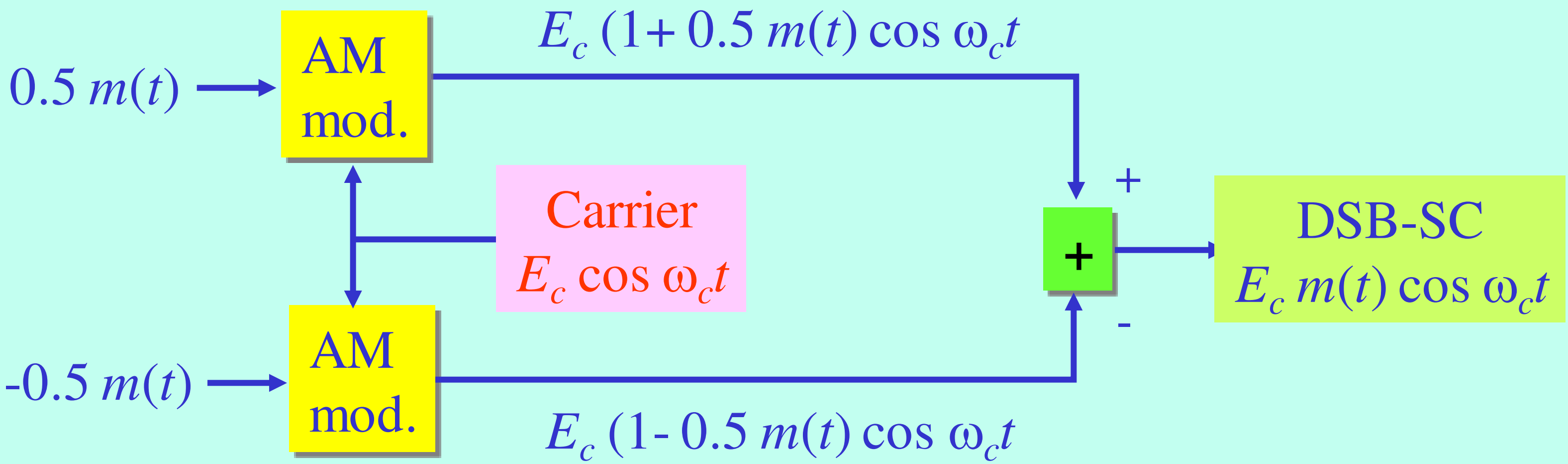


Figure 2.6 DSB-SC amplitude modulation

Notice: No carrier frequency

DSB-SC - *Implementation*

- **Balanced modulator**



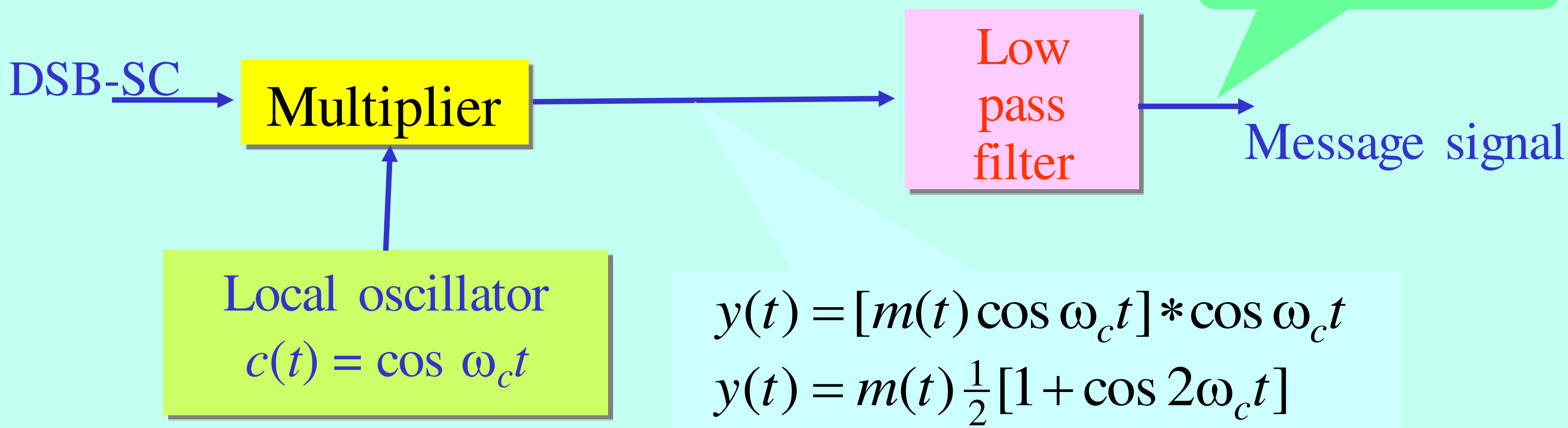
- **Ring modulator**

- **Square-law modulator**

DSB-SC - *Detection*

- **Synchronous detection**

$$v(t) = \tfrac{1}{2} m(t)$$



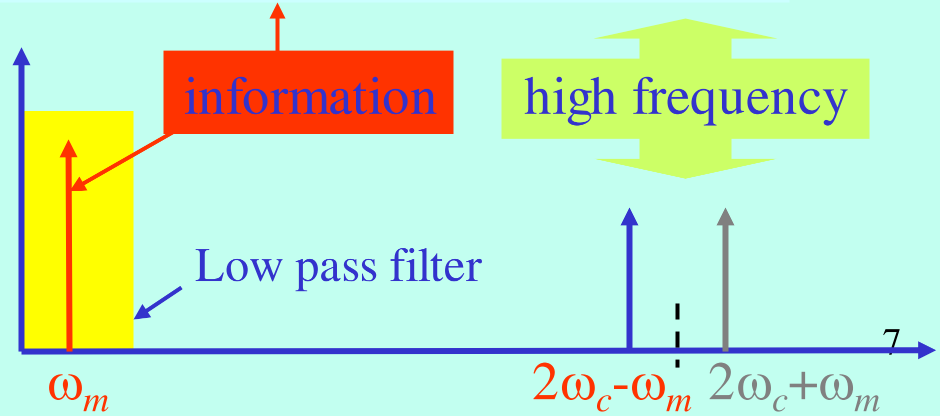
$$y(t) = [m(t)\cos\omega_c t] * \cos\omega_c t$$

$$y(t) = m(t)\tfrac{1}{2}[1+\cos 2\omega_c t]$$

$$= \tfrac{1}{2} m(t) + \tfrac{1}{2} m(t)\cos 2\omega_c t$$

Condition:

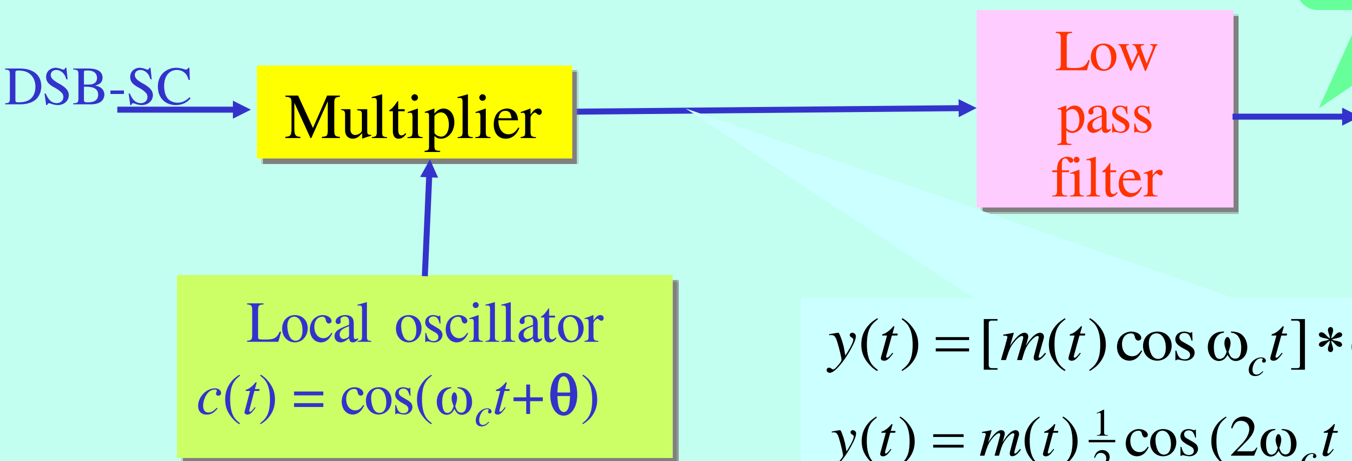
- Local oscillator has the same **frequency** and **phase** as that of the carrier signal at the transmitter.



DSB-SC - *Synchronous Detection*

- **Case 1 - Phase error**

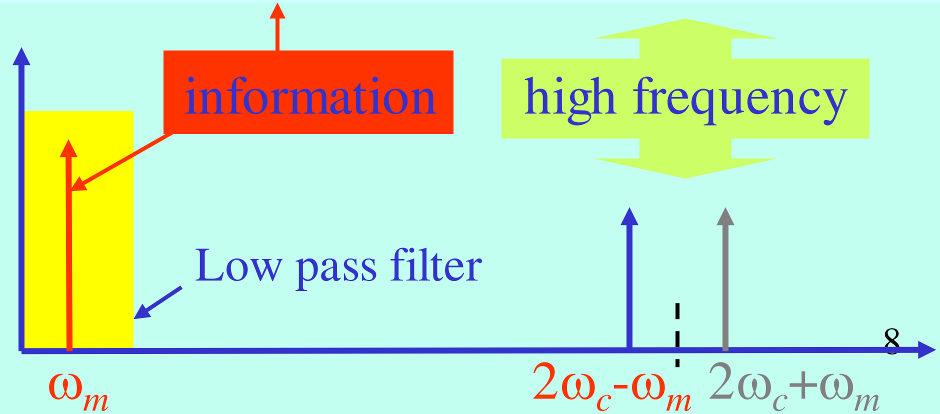
$v(t) = \frac{1}{2} m(t) \cos\theta$



$$y(t) = [m(t)\cos\omega_c t] * \cos(\omega_c t + \theta)$$

$$y(t) = m(t)\tfrac{1}{2}\cos(2\omega_c t + \theta) + m(t)\tfrac{1}{2}\cos(-\theta)$$

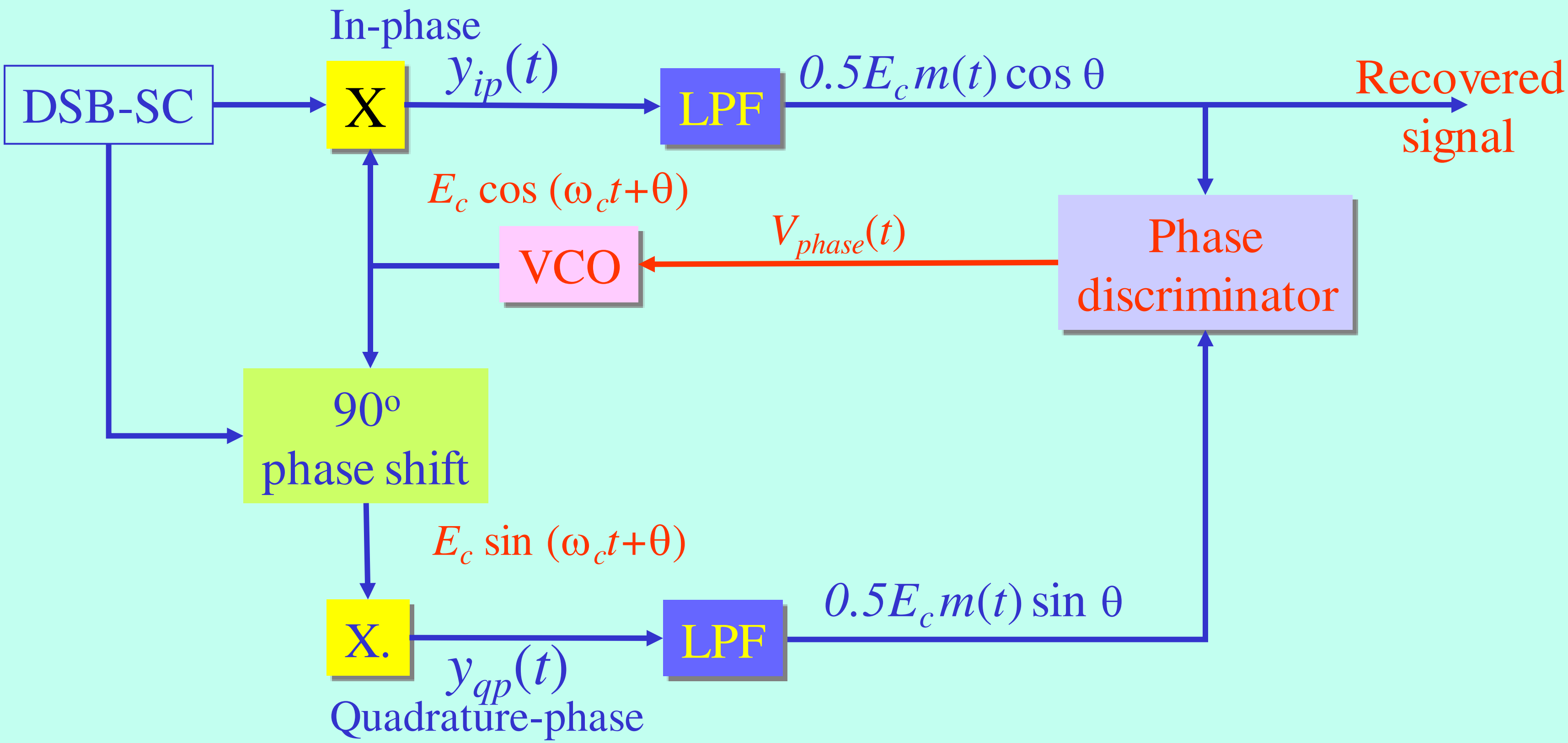
$$= m(t)\tfrac{1}{2}\cos\theta + m(t)\tfrac{1}{2}\cos(2\omega_c t + \theta)$$



Condition:

- Local oscillator has the same **frequency** but *different* **phase** compared to carrier signal at the transmitter.

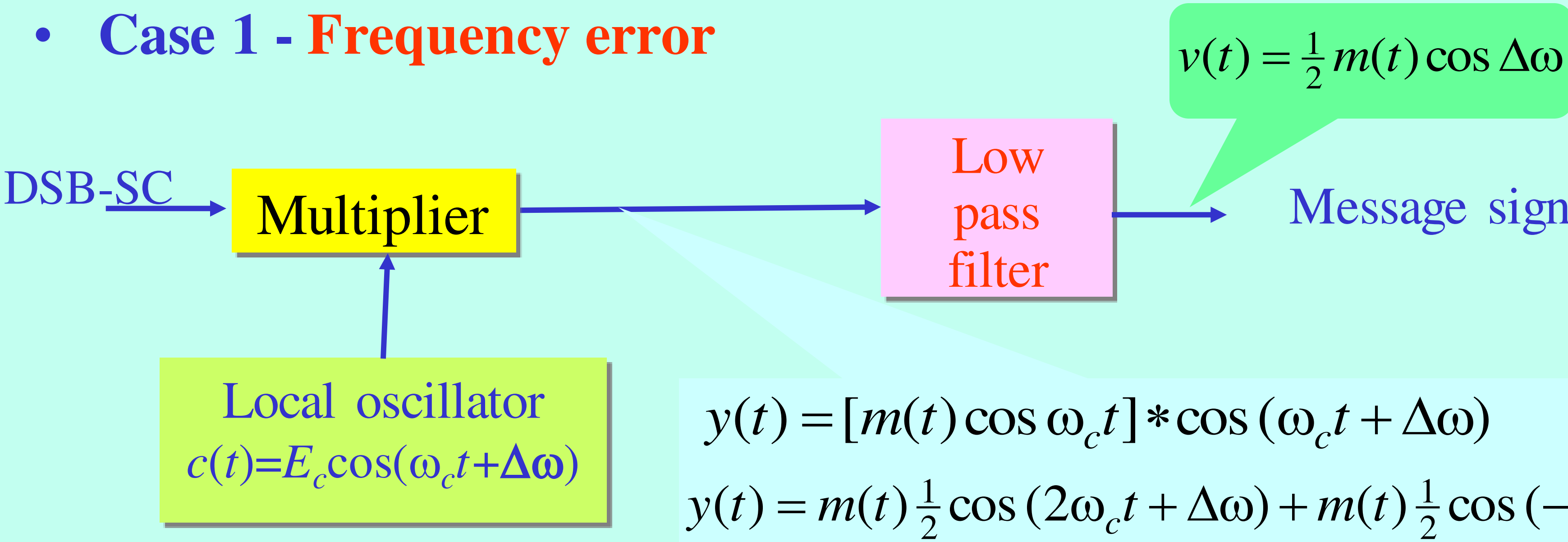
Phase Synchronisation - *Costas Loop*



- When there is no phase error. The quadrature component is zero
- When $\theta \neq 0$, $y_{ip}(t)$ decreases, while $y_{qp}(t)$ increases
- The out put of the phase discriminator is proportional to θ

DSB-SC - *Synchronous Detection*

- **Case 1 - Frequency error**



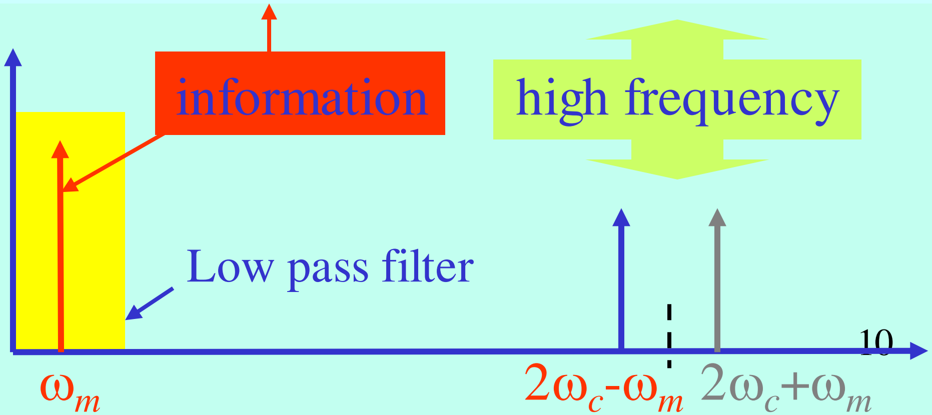
$$y(t)=[m(t)\cos\omega_c t]*\cos(\omega_c t+\Delta\omega)$$

$$y(t)=m(t)\tfrac{1}{2}\cos(2\omega_c t+\Delta\omega)+m(t)\tfrac{1}{2}\cos(-\Delta\omega)$$

$$=m(t)\tfrac{1}{2}\cos\Delta\omega+m(t)\tfrac{1}{2}\cos(2\omega_c t+\Delta\omega)$$

Condition:

•Local oscillator has the same **phase** but *different* **frequency** compared to carrier signal at the transmitter.



DSB-SC - *Power*

- The total power (or average power):

$$P_{T-DSB-SC} = \frac{2}{R}\left[\frac{ME_c/\sqrt{2}}{2}\right]^2 = \frac{(ME_c)^2}{4R}$$

- The maximum and peak envelop power

$$P_{P-DSB-SC} = \frac{(ME_c)^2}{R}$$

DSB-SC - *Summary*

- **Advantages:**
  – Lower power consumption

- **Disadvantage:**
  - Complex detection

- **Applications:**
  - Analogue TV systems: to transmit colour information
  - For transmitting *stereo* information in FM sound broadcast at VHF

Department of ECE
Sub: Analog Communications
Unit - 2
Single Side Band Suppressed Carrier

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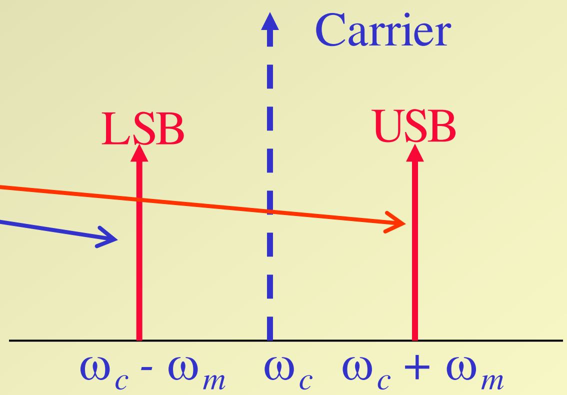
Content

- Theory
- Transmitter Implementation
- Detector Implementation
- Power Analysis
- Summary

Single Side Band Suppressed Carrier

From DSB-SC spectrum:

- Information ω_m is carried twice
- Bandwidth is is high



Single frequency

Question: *Why transmit both side bands?*

Ans:

Question: *Can one suppress one of the side bandcarrier?*

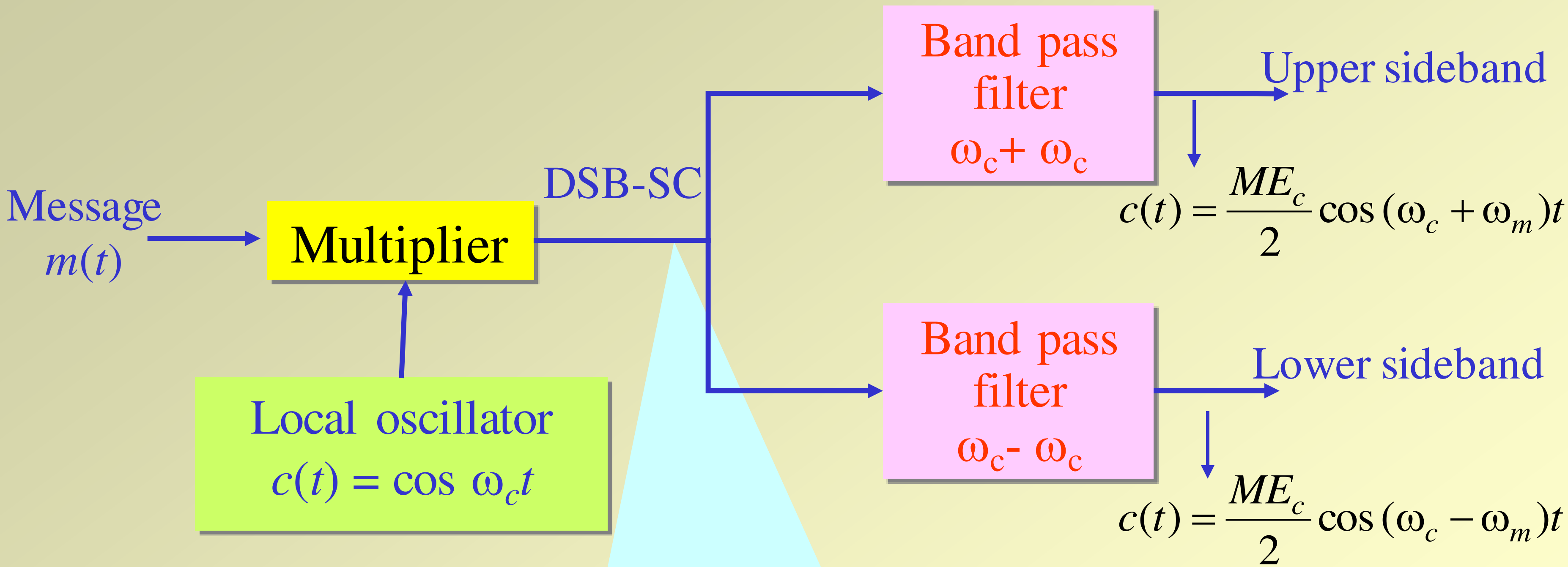
Ans.: *Yes, just transmit one side band (i.e* ***SSB-SC****)*

But what is the penalty?

System complexity at the receiver

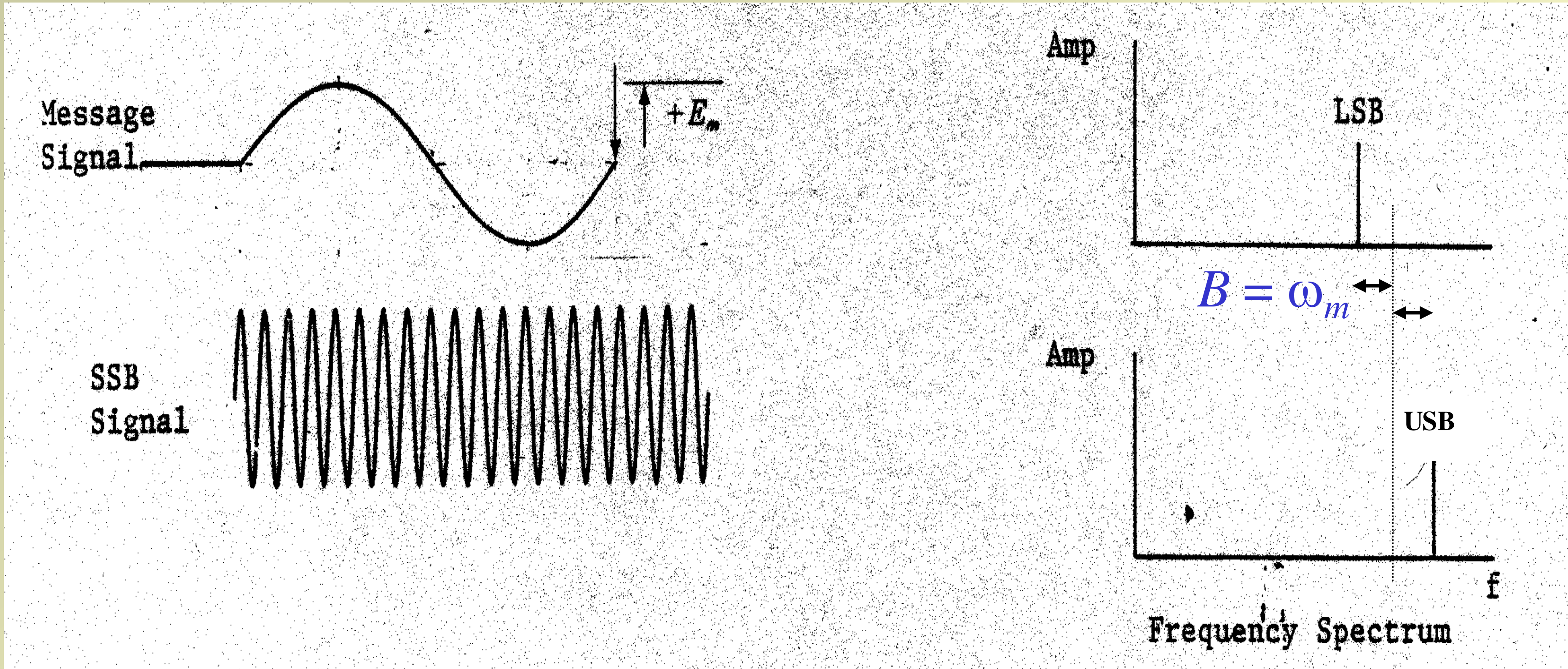
SSB-SC - *Implementation*

- **Frequency discrimination**



$$c(t) = E_m \cos \omega_m t \ \cos \omega_c t$$
$$= \frac{ME_c}{2} \cos(\omega_c + \omega_m)t + \frac{ME_c}{2} \cos(\omega_c - \omega_m)t$$

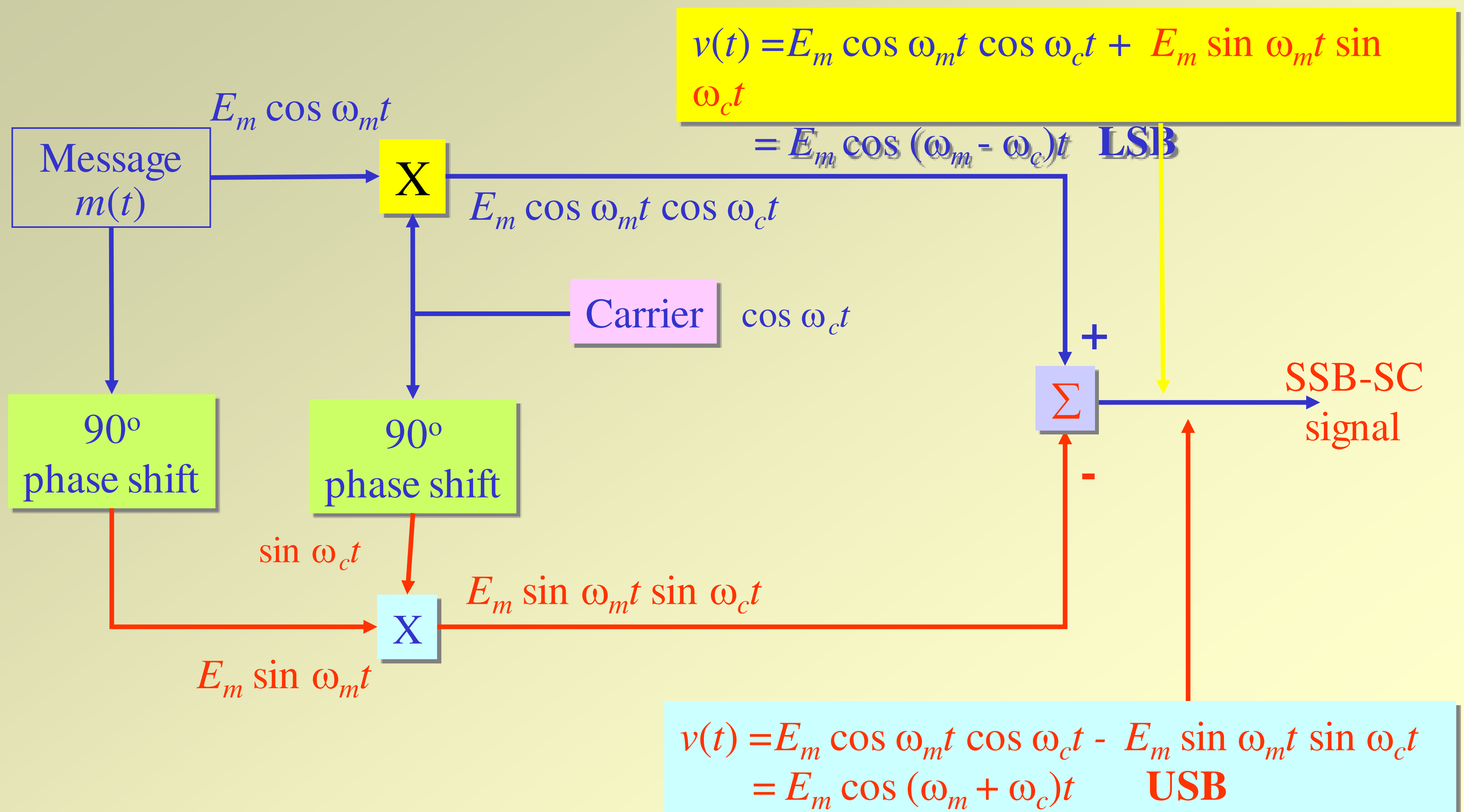
SSB-SC - *Waveforms*



Bandwidth $B = \omega_m$

SSB-SC - *Implementation cont.*

• Phase discrimination (*Hartley modulator*)

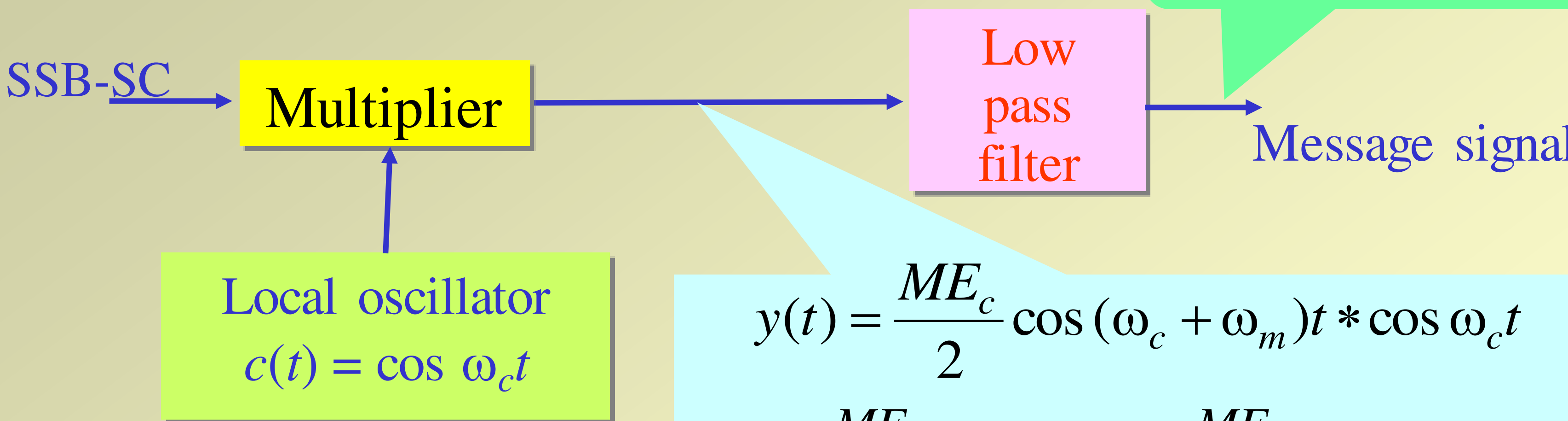


SSB-SC - *Hartley Modulator*

- **Advantages:**
  - No need for bulky and expensive band pass filters
  - Easy to switch from a LSB to an USB SSB output
- **Disadvantage:**
  - Requires Hilbert transform of the message signal. Hilbert transform changes the phase of each +ve frequency component by exactly - 90^o.

SSB-SC - *Detection*

- **Synchronous detection**



$$v(t) = \frac{ME_c}{4}\cos\,\omega_m t$$

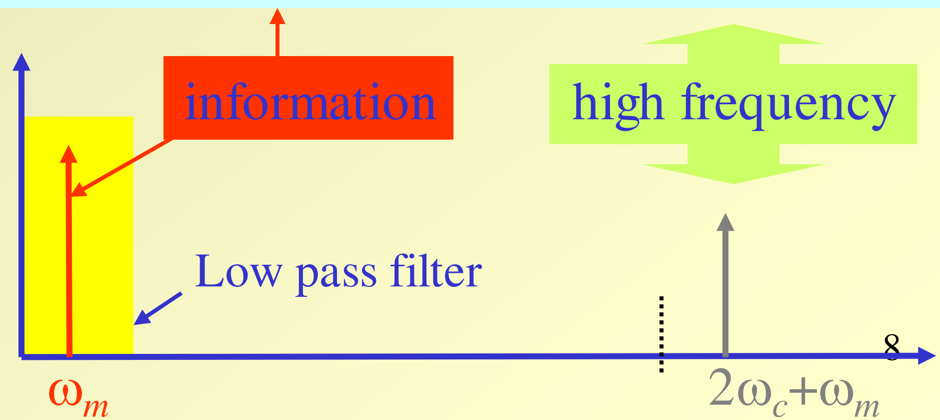
Local oscillator $c(t) = \cos\,\omega_c t$

$$y(t) = \frac{ME_c}{2}\cos(\omega_c + \omega_m)t * \cos\omega_c t$$

$$y(t) = \frac{ME_c}{4}\cos(-\omega_m t) + \frac{ME_c}{4}\cos(2\omega_c + \omega_m)t$$

Condition:

•Local oscillator has the same **frequency** and **phase** as that of the carrier signal at the transmitter.



SSB-SC - *Synch. Detection cont.*

- **Case 1 - Phase error**

SSB-SC → Multiplier → Low pass filter → Message signal

Local oscillator: $c(t) = \cos(\omega_c t + \theta)$ → Multiplier

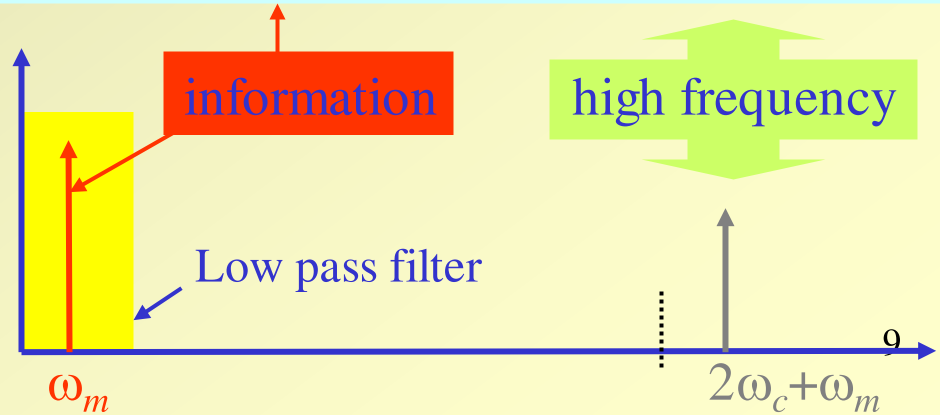
$$v(t) = \frac{ME_c}{4}\cos(\omega_m - \theta)t$$

$$y(t) = \frac{ME_c}{2}\cos(\omega_c + \omega_m)t * \cos(\omega_c t + \theta)$$

$$y(t) = \frac{ME_c}{4}\cos(\omega_m t - \theta) + \frac{ME_c}{4}\cos(2\omega_c t + \omega_m + \theta)$$

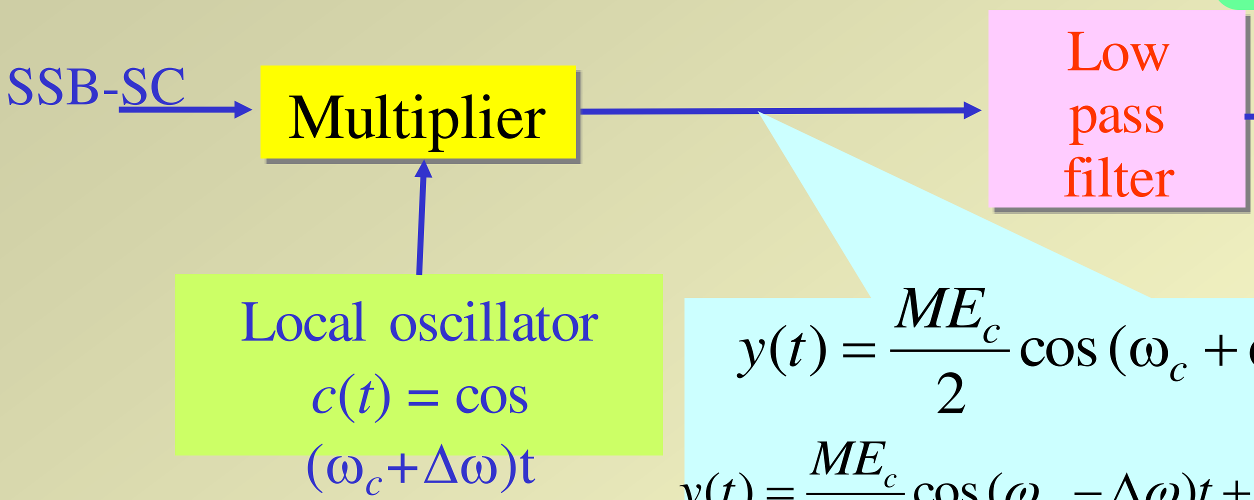
Condition:

- Local oscillator has the same **frequency** but *different* **phase** as that of the carrier signal at the transmitter.



SSB-SC - *Synch. Detection cont.*

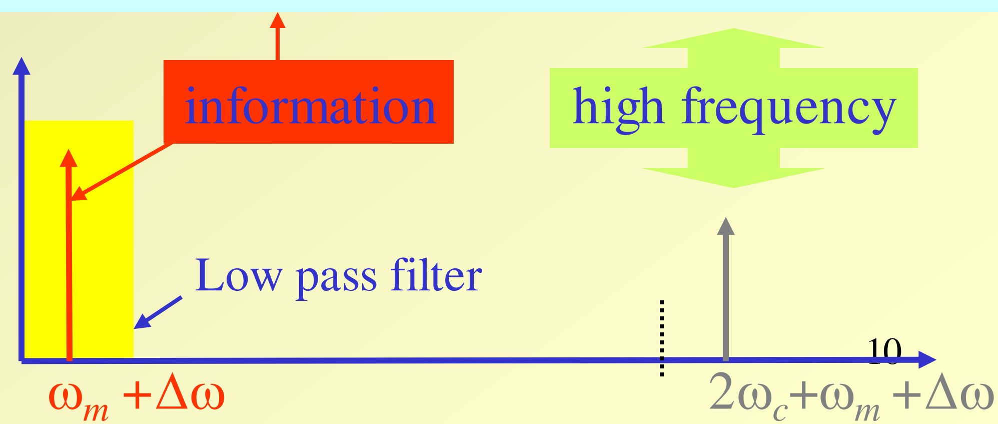
- **Case 1 - Frequency error**



$$v(t) = \frac{ME_c}{4}\cos(\omega_m - \Delta\omega)t$$

$$y(t) = \frac{ME_c}{2}\cos(\omega_c + \omega_m)t * \cos(\omega_c + \Delta\omega)t$$

$$y(t) = \frac{ME_c}{4}\cos(\omega_m - \Delta\omega)t + \frac{ME_c}{4}\cos(2\omega_c + \omega_m + \Delta\omega)t$$



Condition:

- Local oscillator has the same **phase** but *different* **frequency** as that of the carrier signal at the transmitter.

SSB-SC - *Power*

- The total power (or average power):

$$P_{T-SSB-SC} = \frac{1}{R}\left[\frac{ME_c/\sqrt{2}}{2}\right]^2 = \frac{(ME_c)^2}{8R}$$

- The maximum and peak envelop power

$$P_{P-SSB-SC} = \frac{(ME_c)^2}{4R}$$

SSB-SC - *Summary*

- **Advantages:**
  - Lower power consumption
  - Better management of the frequency spectrum
  - Less prone to selective fading
  - Lower noise
- **Disadvantage:**
  - Complex detection
- **Applications:**
  - Two way radio communications
  - Frequency division multiplexing
  - Up conversion in numerous telecommunication systems

Department of ECE
Sub: Analog Communicatons

Unit – 3

FREQUENCY MODULATION AND DEMODULATION

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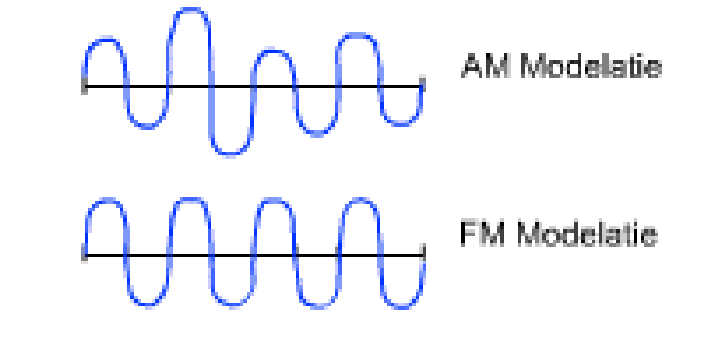
- CONTENT:
  - Angle modulation vs. Amplitude modulation
  - FM Basics
  - Phase Modulation
  - Frequency Modulation
  - FM Characteristics
  - Relationship b/w FM & PM
  - Frequency Modulation
  - Narrow Band FM
  - WideBand FM
  - Bandwidth of FM
  - Phase Locked Loop
  - Advanatages, Disadvantages & Applications of FM

Angle Modulation vs. AM

- Summarize: properties of amplitude modulation
  - Amplitude modulation is linear
    - *just move to new frequency band, spectrum shape does not change. No new frequencies generated.*
  - Spectrum: S(f) is a translated version of M(f)
  - Bandwidth $\le$ 2W
- Properties of angle modulation
  - They are nonlinear
    - *spectrum shape does change, new frequencies generated.*
  - S(f) is not just a translated version of M(f)
  - Bandwidth is usually much larger than 2W

FM Basics

- VHF (30M-300M) high-fidelity broadcast
- Wideband FM, (FM TV), narrow band FM (two-way radio)
- 1933 **FM** and angle modulation proposed by Armstrong, but success by 1949.
- Digital: Frequency Shift Key (FSK), Phase Shift Key (BPSK, QPSK, 8PSK,…)
- AM/FM: Transverse wave/Longitudinal wave



Instantaneous Frequency

- Angle modulation has two forms
  - Frequency modulation (FM): message is represented as the variation of the instantaneous frequency of a carrier
  - Phase modulation (PM): message is represented as the variation of the instantaneous phase of a carrier

$$s(t) = A_c \cos\left[\theta_i(t)\right],$$

where A_c : carrier amplitude, $\theta_i(t)$: angle (phase)

$$f_i(t) = \frac{1}{2\pi}\frac{d\theta_i(t)}{dt}$$

$$s(t) = A_c \cos\left[2\pi f_c t + \phi(t)\right]$$

where $\phi(t)$ is a function of message signal $m(t)$.

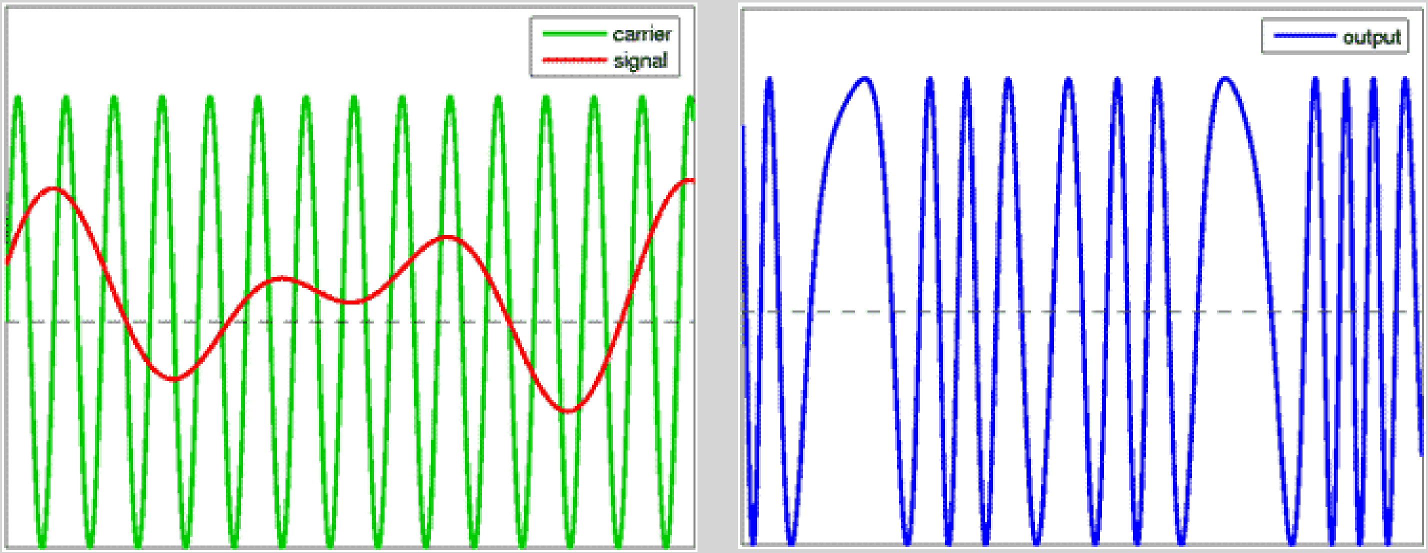
Phase Modulation

- PM (phase modulation) signal

$$s(t) = A_c \cos\left[2\pi f_c t + k_p m(t)\right]$$

$$\phi(t) = k_p m(t), \quad k_p : \text{phase sensitivity}$$

$$\text{instantanous frequency } f_i(t) = f_c + \frac{k_p}{2\pi}\frac{dm(t)}{dt}$$



Frequency Modulation

- FM (frequency modulation) signal

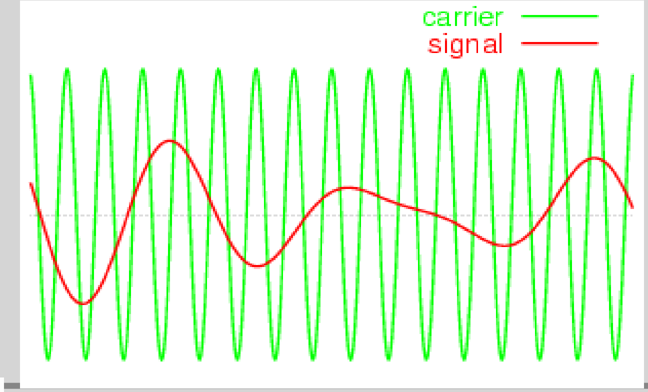
$$s(t) = A_c \cos\left[2\pi f_c t + 2\pi k_f \int_0^t m(\tau)d\tau\right]$$

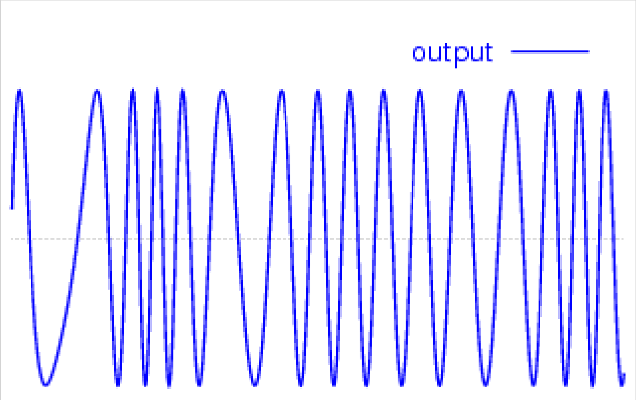
k_f : frequency sensitivity

instantanous frequency $f_i(t) = f_c + k_f m(t)$

angle $\theta_i(t) = 2\pi \int_0^t f_i(\tau)d\tau$ (Assume zero initial phase)

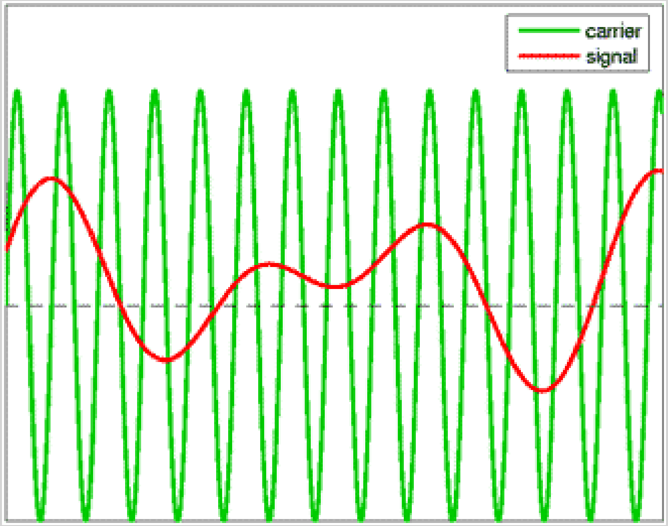
$$= 2\pi f_c t + 2\pi k_f \int_0^t m(\tau)d\tau$$

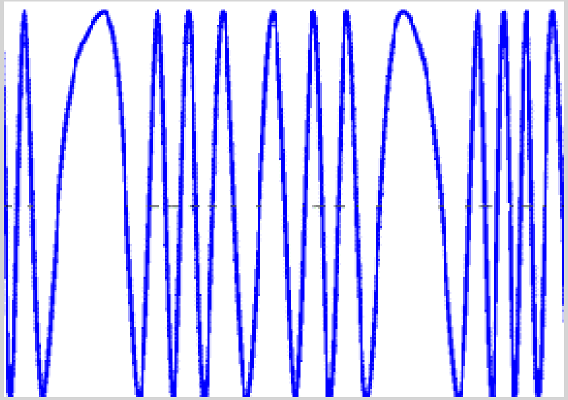


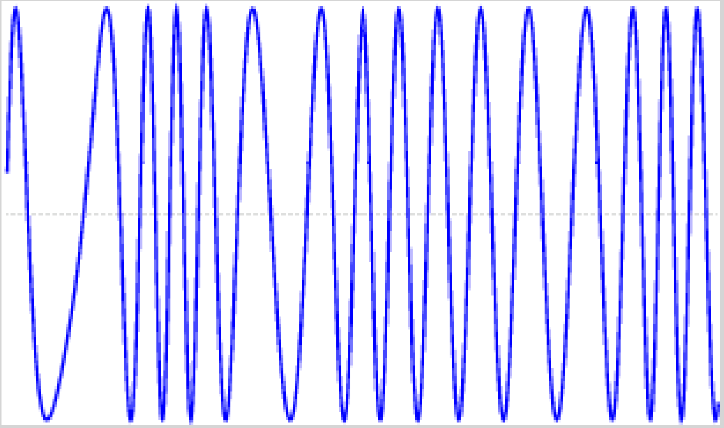


FM Characteristics

- Characteristics of FM signals
  - Zero-crossings are not regular
  - Envelope is constant
  - FM and PM signals are similar

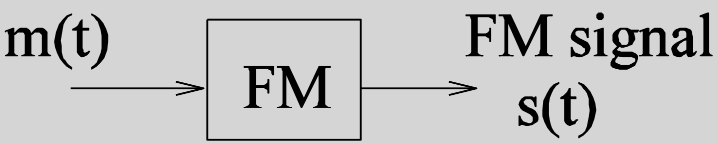


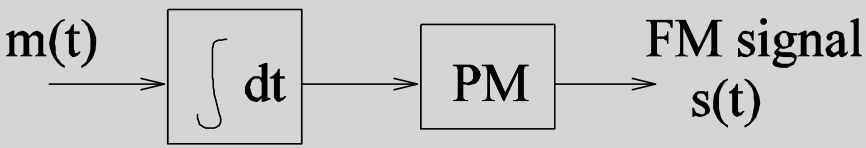




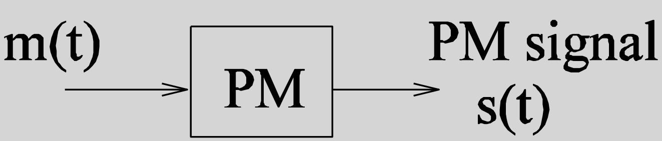
Relations between FM and PM

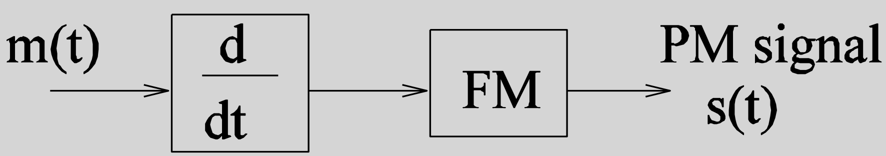
$$\text{FM of } m(t) \;\Leftrightarrow\; \text{PM of } \int_0^t m(\tau)\,d\tau$$



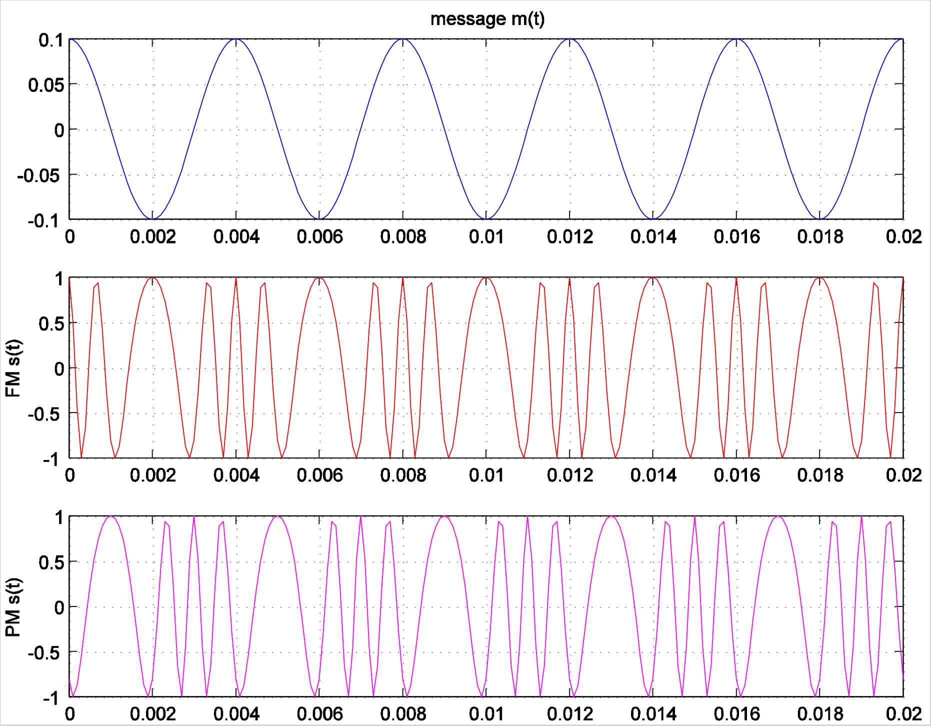


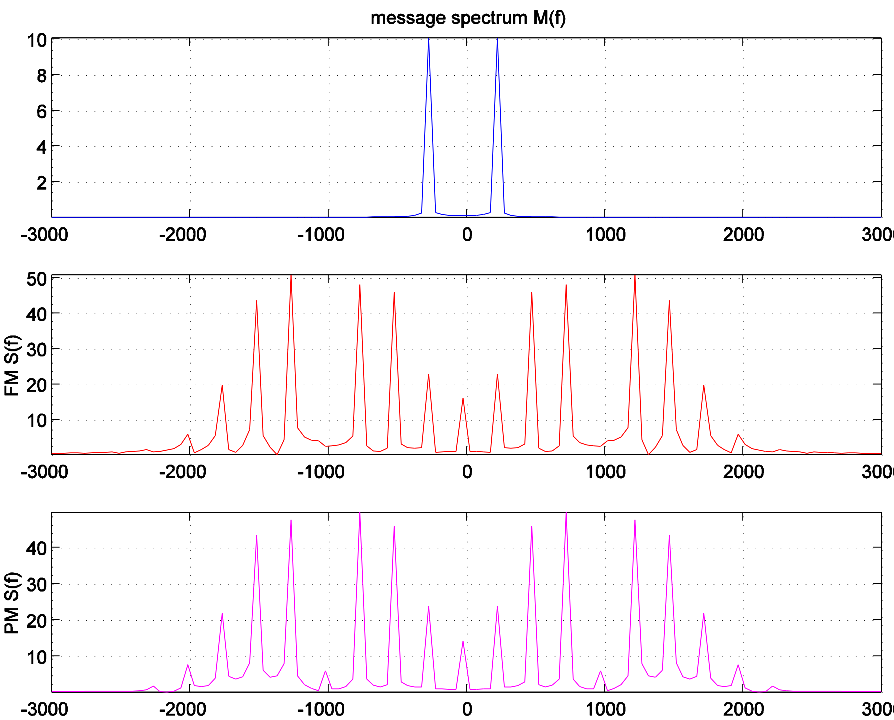
$$\text{PM of } m(t) \;\Leftrightarrow\; \text{FM of } \frac{dm(t)}{dt}$$





FM/PM Example (Time/Frequency)





Frequency Modulation

- FM (frequency modulation) signal

$$s(t) = A_c \cos\left[2\pi f_c t + 2\pi k_f \int_0^t m(\tau) d\tau \right]$$

k_f : frequency sensitivity

instantanous frequency $f_i(t) = f_c + k_f m(t)$

angle $\theta_i(t) = 2\pi \int_0^t f_i(\tau) d\tau$ (Assume zero initial phase)

$$= 2\pi f_c t + 2\pi k_f \int_0^t m(\tau) d\tau$$

$m(t) = A_m \cos(2\pi f_m t)$ $f_i = f_c + k_f A_m \cos(2\pi f_m t)$

$$f_i = \frac{1}{2\pi} \frac{d\theta}{dt} = \frac{1}{2\pi} \frac{d\left[2\pi f_c t\right]}{dt} + \frac{1}{2\pi} \frac{d\left[2\pi k_f \int_0^t A_m \cos(2\pi f_m \tau) d\tau \right]}{dt}$$

$$= f_c + \frac{1}{2\pi} 2\pi k_f \left[A_m \cos(2\pi f_m \tau) \right]\Big|_{\text{Let } \tau = t}$$

Example

Consider m(t)- a square wave- as shown. The FM wave for this m(t) is shown below.

$$\varphi_{FM}(t) = A\cos(\omega_c t + k_f \int_{-\infty}^{t} m(\lambda)d\lambda).$$

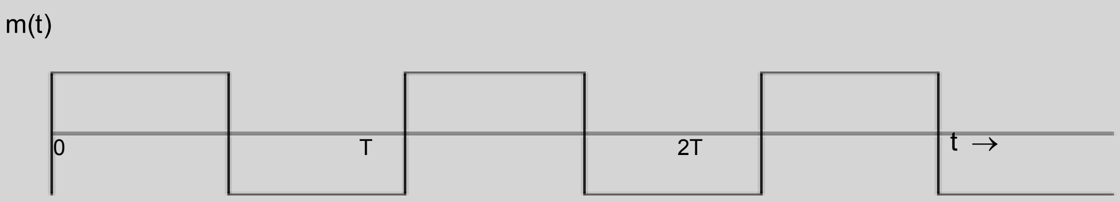
Assume m(t) starts at $t = 0$. For $0 \le t \le \frac{T}{2}$ $\quad m(t) = 1$, $\int_0^t m(\lambda)d\lambda = t$ and

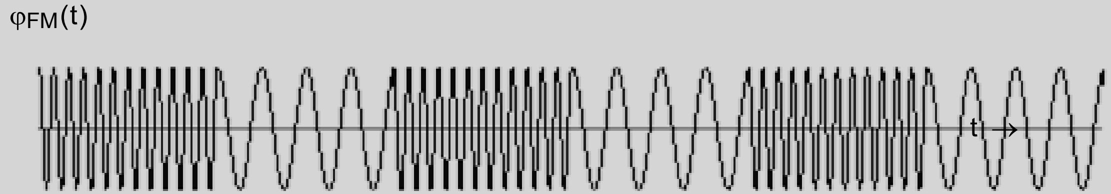
for $\frac{T}{2} \le t \le T$ $\quad m(t) = -1$, $\int_0^t m(\lambda)d\lambda = \int_0^{\frac{T}{2}} m(\lambda)d\lambda + \int_{\frac{T}{2}}^{t} m(\lambda)d\lambda = \frac{T}{2} - (t - \frac{T}{2}) = T - t.$

The instantaneous frequency is $\omega_i(t) = \omega_c + k_f\, m(t) = \omega_c + k_f$ for $0 \le t \le \frac{T}{2}$

and $\omega_i(t) = \omega_c - k_f$ for $\frac{T}{2} \le t \le T$.

$\omega_{i\,max} = \omega_c + k_f$ and $\omega_{i\,min} = \omega_c - k_f$





Frequency Deviation

- Frequency deviation Δf
  - difference between the maximum instantaneous and carrier frequency
  - Definition: $\Delta f = k_f A_m = k_f \max|m(t)|$
  - Relationship with instantaneous frequency

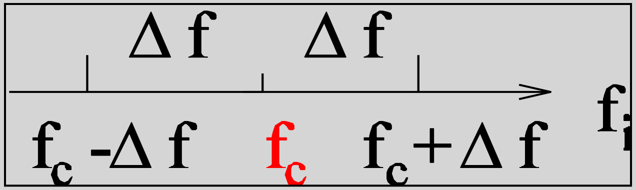
$$\text{single-tone } m(t) \text{ case: } f_i = f_c + \Delta f \cos(2\pi f_m t)$$

$$\text{general case: } f_c - \Delta f \le f_i \le f_c + \Delta f$$

  - Question: Is bandwidth of s(t) just 2Δf?

No, instantaneous frequency is not equivalent to spectrum frequency (with non-zero power)!

S(t) has ∞ spectrum frequency (with non-zero power).



Modulation Index

- Indicate by how much the modulated variable (instantaneous frequency) varies around its unmodulated level (message frequency)

$$\text{AM (envelope): } \frac{\max|\ m(t)|}{\mathrm{A}},$$

$$\text{FM (frequency): } \beta = \frac{\max|k_f m(t)|}{f_m}$$

- Bandwidth

$$a(t) = \int_{-\infty}^{t} m(\alpha) d\alpha$$

$$\varphi(t) = \mathrm{Re}(\varphi(t)) = A\left[\cos w_c t - k_f a(t) \sin w_c t - \frac{k_f^2}{2!} a^2(t) \cos w_c t + \frac{k_f^2}{3!} a^3(t) \sin w_c t \ldots\right]$$

Narrow Band Angle Modulation

Definition $\left|k_f a(t)\right| << 1$

Equation $\varphi(t) = A\left[\cos w_c t - k_f a(t) \sin w_c t\right]$

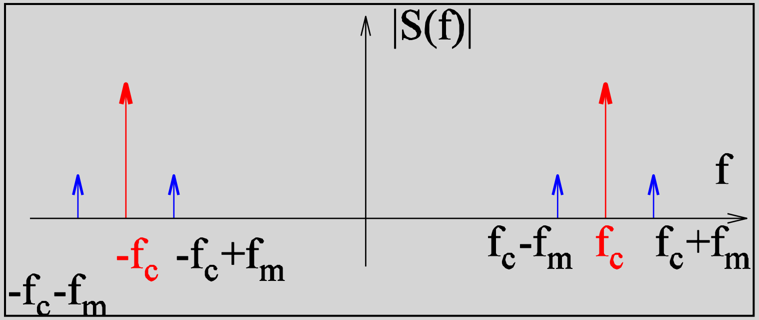
Comparison with AM

Only phase difference of Pi/2

Frequency: similar

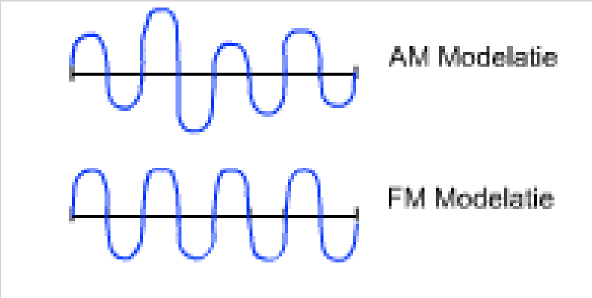
Time: AM: frequency constant

FM: amplitude constant

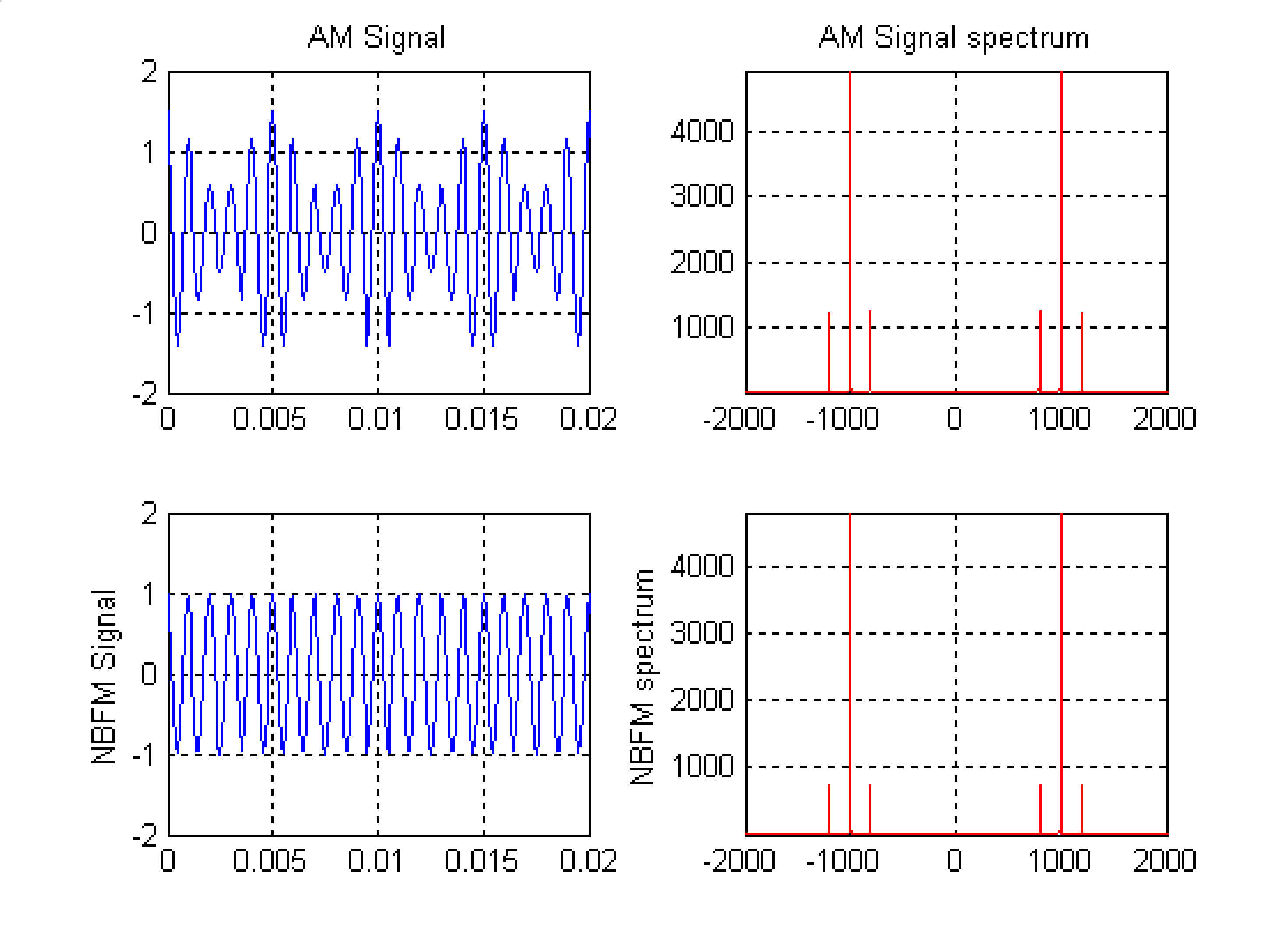


Conclusion: NBFM signal is similar to AM signal

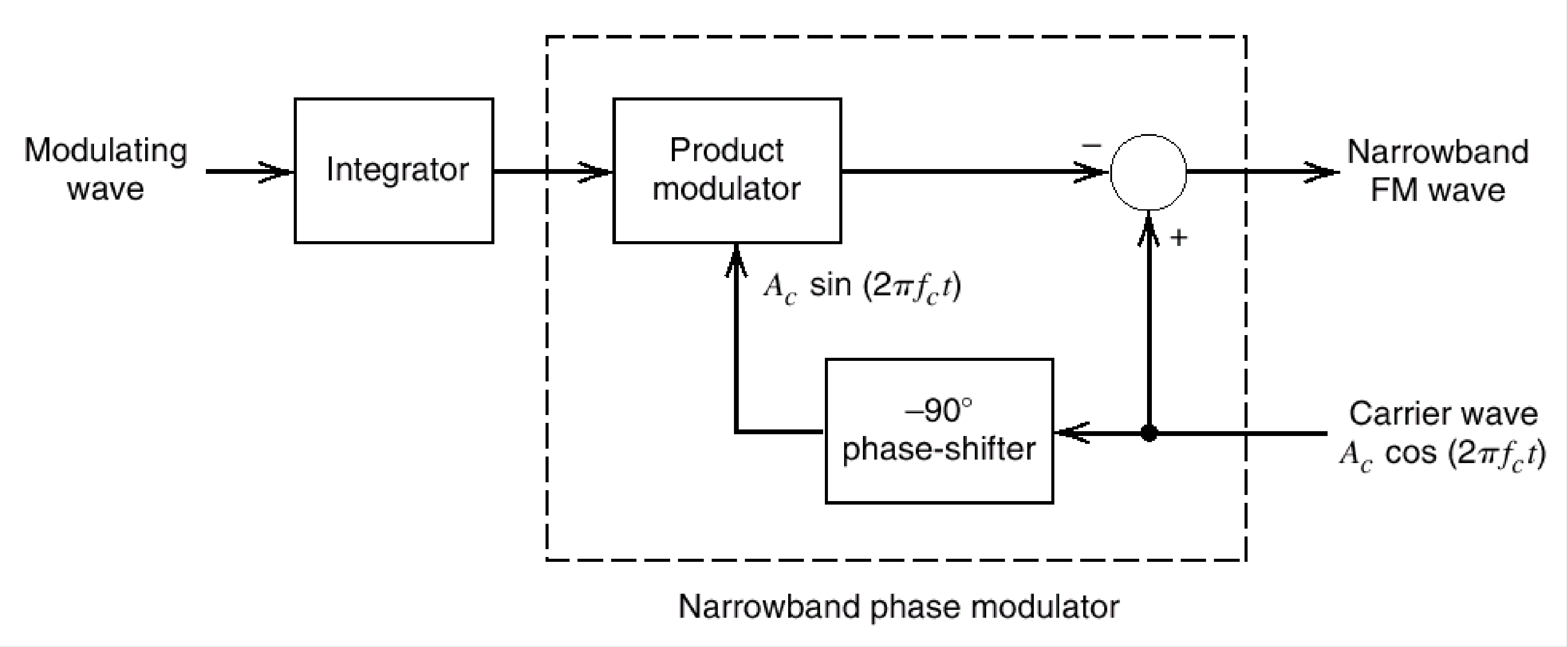
NBFM has also bandwidth 2W. (twice message signal bandwidth)



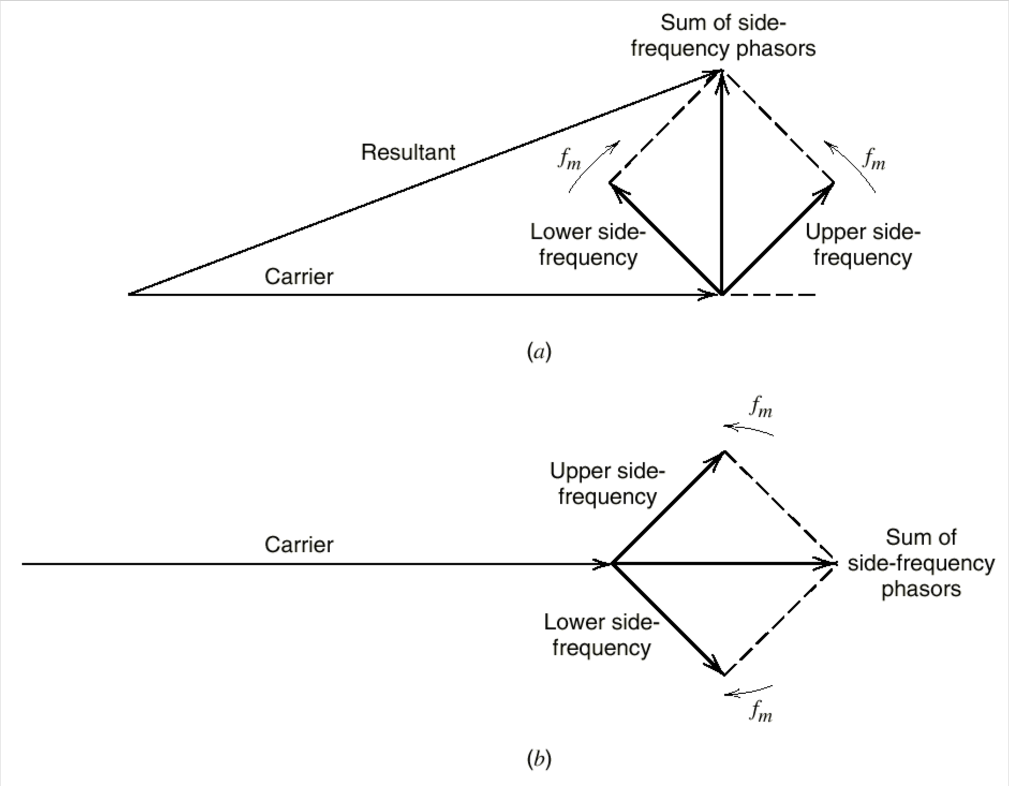
Example


AM Signal
2
1
0
-1
-2
0
0.005
0.01
0.015
0.02
AM Signal spectrum
4000
3000
2000
1000
-2000
-1000
0
1000
2000
NBFM Signal
2
1
0
-1
-2
0
0.005
0.01
0.015
0.02
NBFM spectrum
4000
3000
2000
1000
-2000
-1000
0
1000
2000

Block diagram of a method for generating a narrowband FM signal.



A phasor comparison of narrowband FM and AM waves for sinusoidal modulation. (a) Narrowband FM wave. (b) AM wave.



Wide Band FM

- Wideband FM signal

$$m(t) = A_m \cos(2\pi f_m t)$$

$$s(t) = A_c \cos\left[2\pi f_c t + \beta \sin(2\pi f_m t)\right]$$

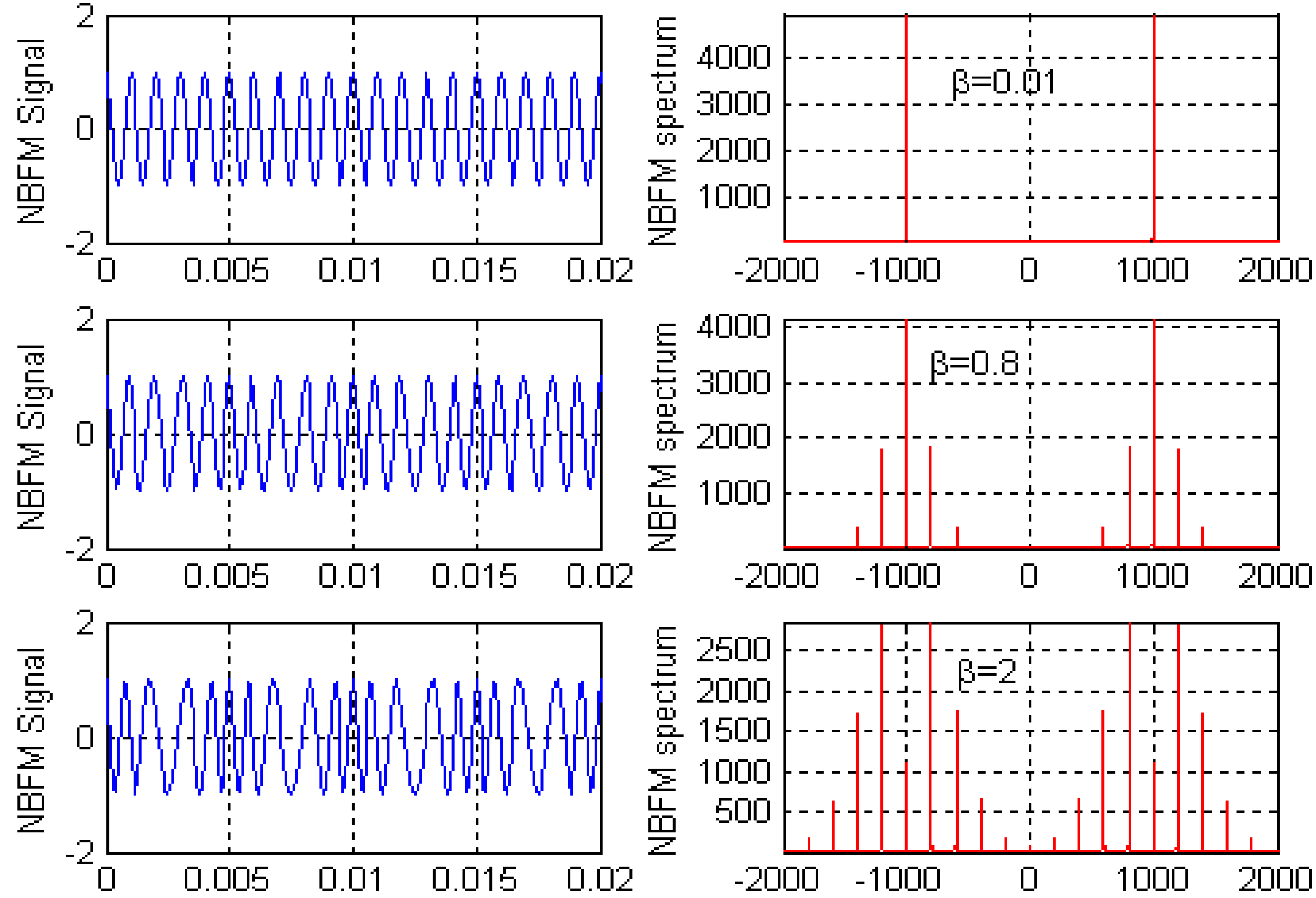
- Fourier series representation

$$s(t) = A_c \sum_{n=-\infty}^{\infty} J_n(\beta) \cos\left[2\pi (f_c + n f_m) t\right]$$

$$S(f) = \frac{A_c}{2} \sum_{n=-\infty}^{\infty} J_n(\beta) \left[\delta(f - f_c - n f_m) + \delta(f + f_c + n f_m)\right]$$

$J_n(\beta)$: n-th order Bessel function of the first kind

Example



Bessel Function of First Kind

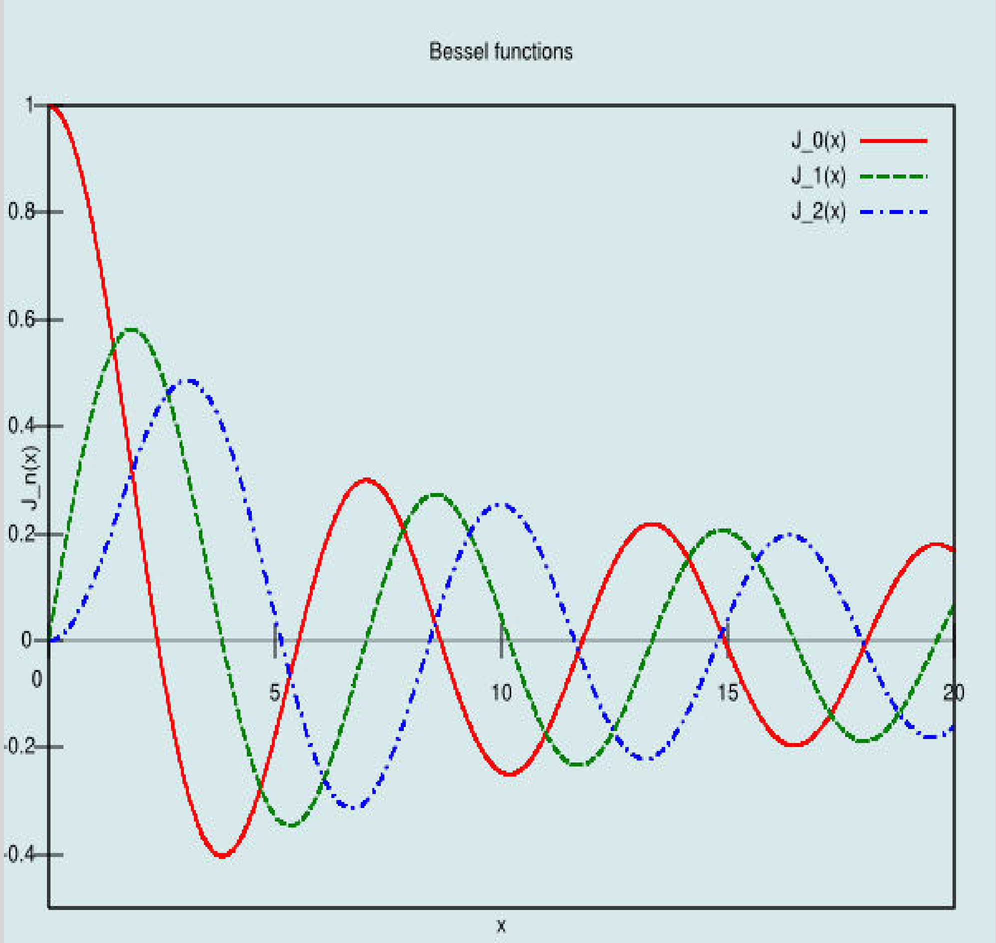
1. $J_n(\beta) = (-1)^n J_{-n}(\beta)$

2. If β is small, then $J_0(\beta) \approx 1,$

$$J_1(\beta) \approx \frac{\beta}{2},$$

$$J_n(\beta) \approx 0 \quad \text{for all } n > 2$$

3. $\sum_{n=-\infty}^{\infty} J_n^2(\beta) = 1$



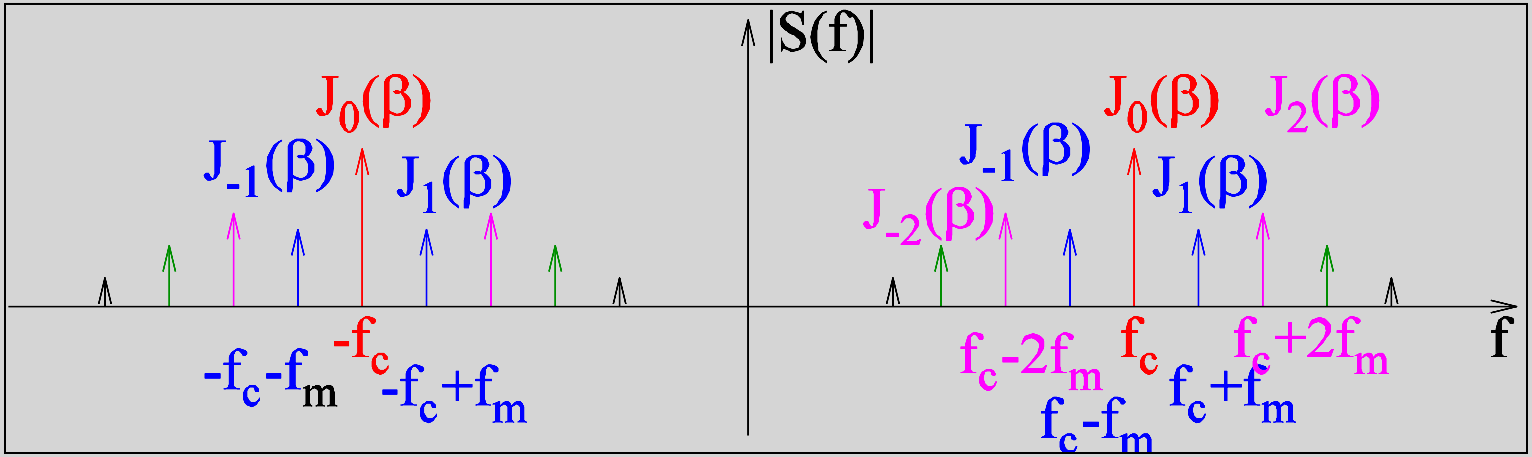
Spectrum of WBFM (Chapter 5.2)

- Spectrum when m(t) is single-tone

$$s(t) = A_c \cos\left[2\pi f_c t + \beta \sin(2\pi f_m t)\right] = A_c \sum_{n=-\infty}^{\infty} J_n(\beta)\cos\left[2\pi(f_c + nf_m)t\right]$$

$$S(f) = \frac{A_c}{2}\sum_{n=-\infty}^{\infty} J_n(\beta)\left[\delta(f - f_c - nf_m) + \delta(f + f_c + nf_m)\right]$$

- Example 2.2



Bandwidth of FM

- Facts
  - FM has side frequencies extending to infinite frequency → theoretically infinite bandwidth
  - But side frequencies become negligibly small beyond a point → practically finite bandwidth
  - FM signal bandwidth equals the required transmission (channel) bandwidth
- Bandwidth of FM signal is approximately by
  - Carson's Rule (which gives lower-bound)

Carson's Rule

- Nearly all power lies within a bandwidth of
  - For single-tone message signal with frequency f_m

$$B_T = 2\Delta f + 2f_m = 2(\beta + 1)f_m$$

  - For general message signal m(t) with bandwidth (or highest frequency) W

$$B_T = 2\Delta f + 2W = 2(D+1)W$$

where $D = \dfrac{\Delta f}{W}$ is deviation ratio (equivalent to β),

$$\Delta f = \max\left[k_f m(t)\right]$$

Phase-Locked Loop

- Can be a whole course. The most important part of receiver.
- Definition: a closed-loop feedback control system that generates and outputs a signal in relation to the frequency and phase of an input ("reference") signal
- A phase-locked loop circuit responds both to the frequency and phase of the input signals, automatically raising or lowering the frequency of a controlled oscillator until it is matched to the reference in both frequency and phase.

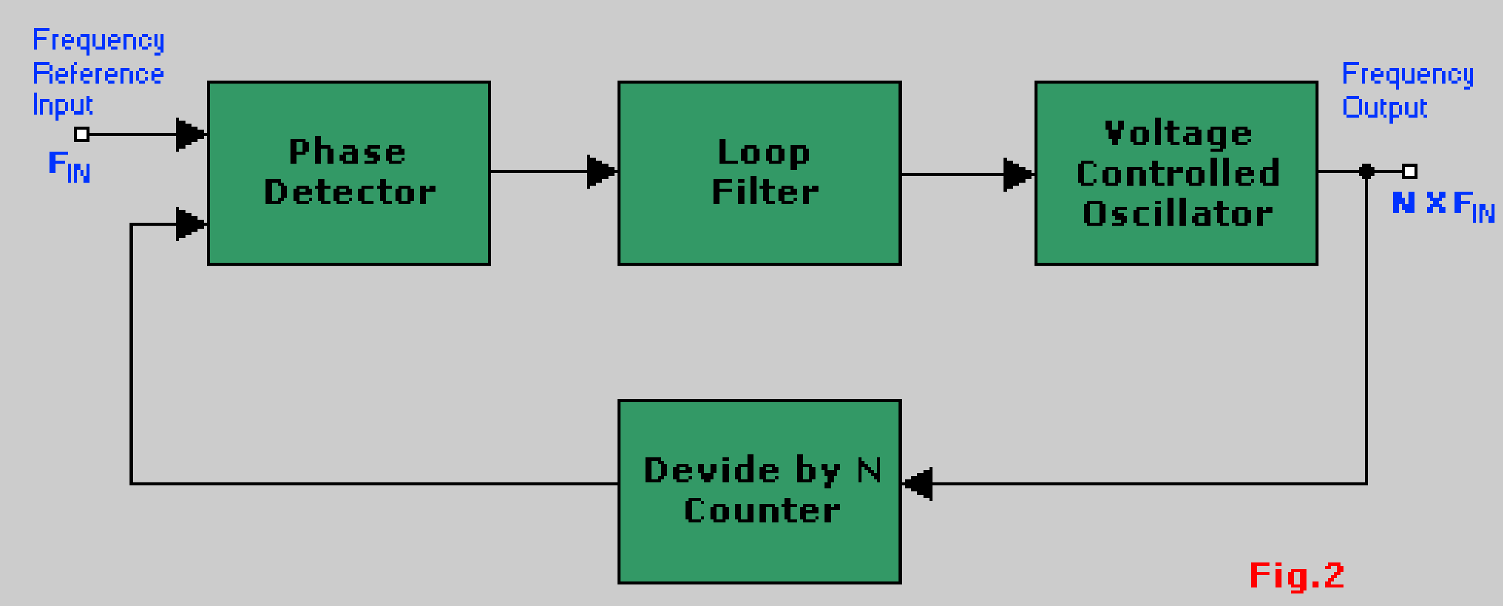
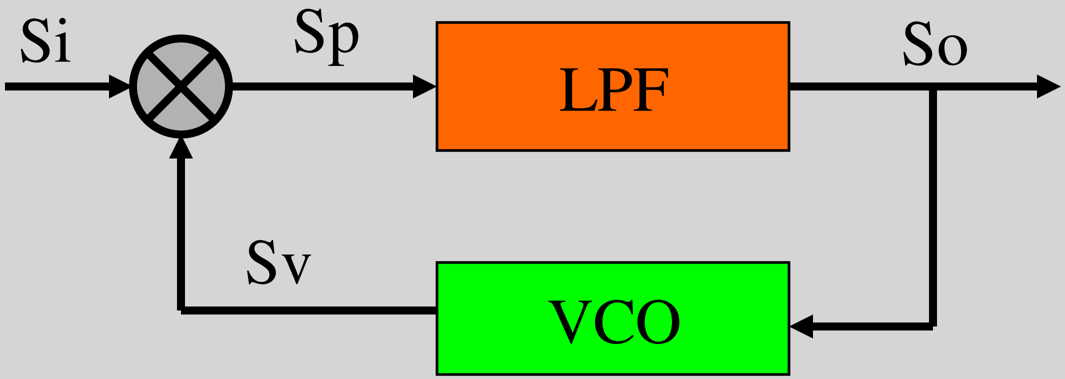


Fig.2

Ideal Model

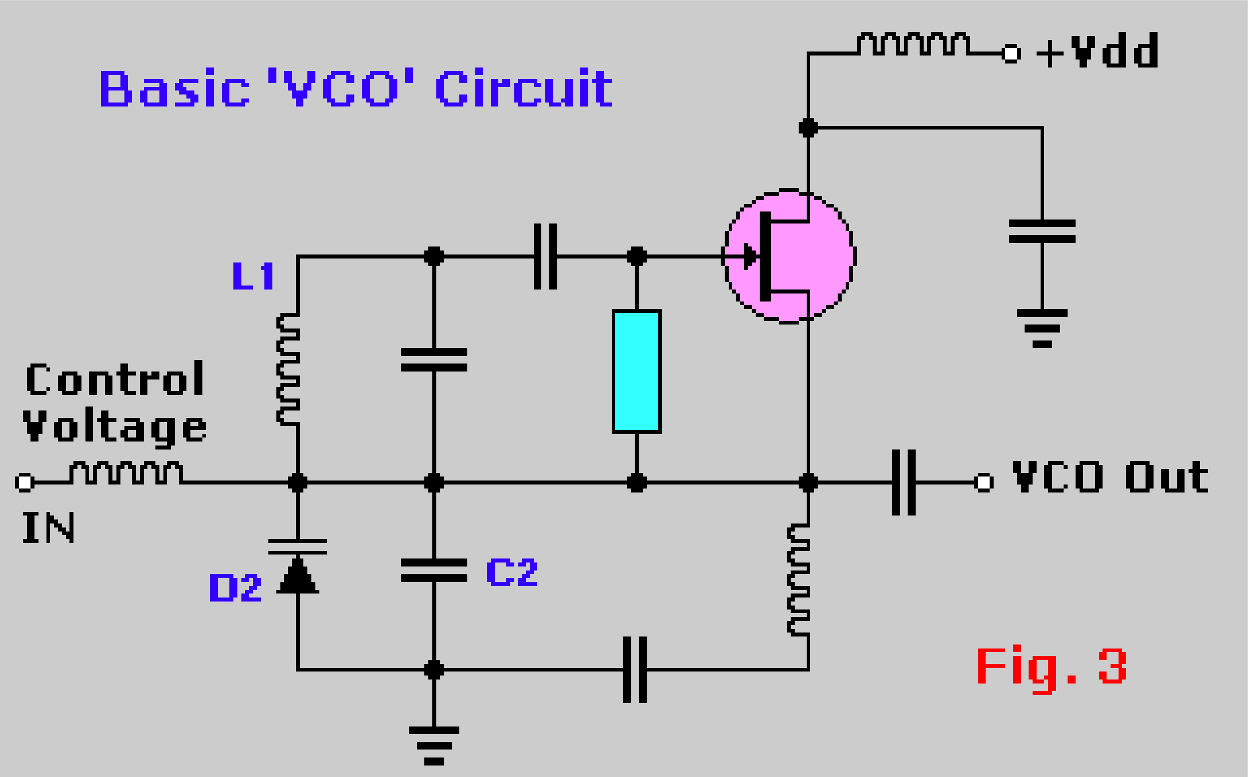
- Model



  - $S_i = A\cos(w_c t + \phi_1(t))$, $S_v = A_v\cos(w_c t + \phi_c(t))$
  - $S_p = 0.5AA_v[\sin(2w_c t + \phi_1 + \phi_c) + \sin(\phi_1 - \phi_c)]$
  - $S_o = 0.5AA_v\sin(\phi_1 - \phi_c) = AA_v(\phi_1 - \phi_c)$
- Capture Range and Lock Range

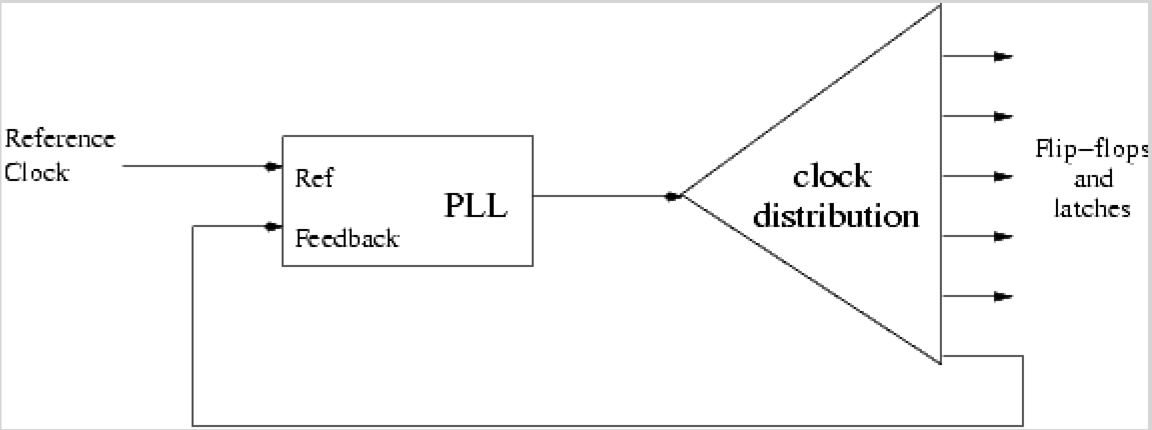
Voltage Controlled Oscillator (VCO)

- $W(t)=w_c+ce_0(t)$, where w_c is the free-running frequency
- Example



PLL Applications

- Clock recovery: no pilot
- Deskewing: circuit design
- Clock generation: Direct Digital Synthesis
- Spread spectrum:
- Jitter Noise Reduction
- Clock distribution



Angle Modulation Pro/Con Application

- Why need angle modulation?
  - Better noise reduction
  - Improved system fidelity
- Disadvantages
  - Low bandwidth efficiency
  - Complex implementations
- Applications
  - FM radio broadcast
  - TV sound signal
  - Two-way mobile radio
  - Cellular radio
  - Microwave and satellite communications

Department of ECE
Sub: Analog Communications
Unit - 4
Effect of Noise on Analog Communication Systems

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AVNIET

Content:

- **Effect of Noise on a Baseband System**
- **Effect of Noise on DSB-SC AM**
- **Effect of Noise on SSB-AM**
- **Effect of Noise on Conventional AM**

Introduction

- Angle modulation systems and FM can provide a high degree of noise immunity
- This noise immunity is obtained at the price of sacrificing channel bandwidth
- Bandwidth requirements of angle modulation systems are considerably higher than that of amplitude modulation systems
- This chapter deals with the followings:
  - Effect of noise on amplitude modulation systems
  - Effect of noise on angle modulation systems
  - Carrier-phase estimation using a phase-locked loop (PLL)
  - Analyze the effects of transmission loss and noise on analog communication systems

Effect of Noise on a Baseband System

- Since baseband systems serve as a basis for comparison of various modulation systems, we begin with a noise analysis of a baseband system.
- In this case, there is no carrier demodulation to be performed.
- The receiver consists only of an ideal lowpass filter with the bandwidth W.
- The noise power at the output of the receiver, for a white noise input, is

$$P_{n_0} = \int_{-W}^{W} \frac{N_0}{2}\, df = N_0 W$$

- If we denote the received power by P_R, the baseband SNR is given by

$$\left(\frac{S}{N}\right)_b = \frac{P_R}{N_0 W} \qquad (6.1.2)$$

White process

- *White* process is processes in which all frequency components appear with equal power, i.e., the power spectral density (PSD), $S_x(f)$, is a constant for all frequencies.
- the PSD of thermal noise, $S_n(f)$, is usually given as $S_n(f) = \frac{kT}{2}$ (where k is Boltzrnann's constant *and* T is the temperature)
- The value kT is usually denoted by N_0, *Then* $S_n(f) = \frac{N_0}{2}$

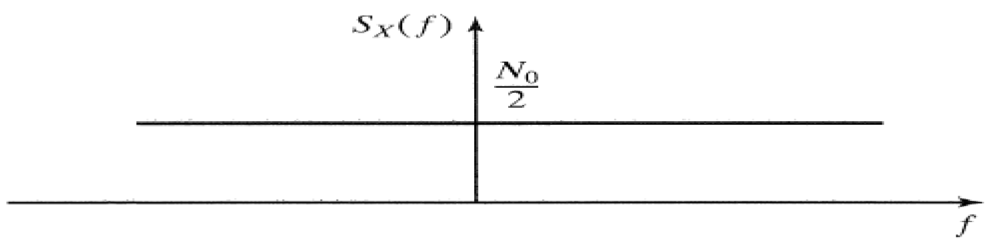

Figure 5.19 Power spectrum of a white process.

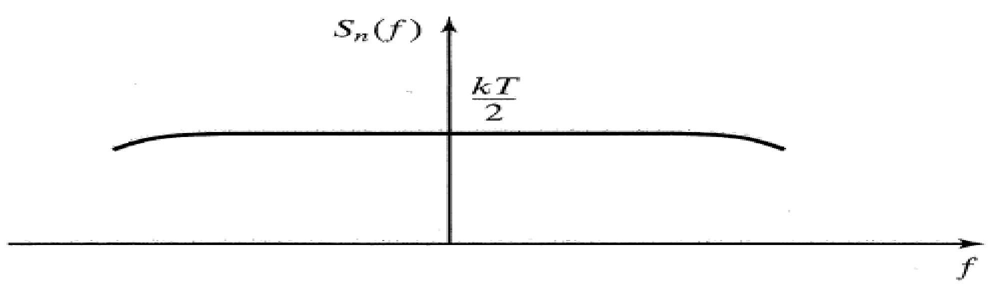

Figure 5.20 Power spectrum of thermal noise.

Effect of Noise on DSB-SC AM

- Transmitted signal : $u(t) = A_c m(t)\cos(2\pi f_c t)$
- The received signal at the output of the receiver noise-limiting filter : Sum of this signal and filtered noise
- Recall from Section 5.3.3 and 2.7 that a filtered noise process can be expressed in terms of its in-phase and quadrature components as

$$n(t) = A(t)\cos[2\pi f_c t + \theta(t)] = A(t)\cos\theta(t)\cos(2\pi f_c t) - A(t)\sin\theta(t)\sin(2\pi f_c t)$$
$$= n_c(t)\cos(2\pi f_c t) - n_s(t)\sin(2\pi f_c t)$$

(where $n_c(t)$ is in-phase component and $n_s(t)$ is quadrature component)

Effect of Noise on DSB-SC AM

- Received signal (Adding the filtered noise to the modulated signal)

$$
\begin{aligned}
r(t) &= u(t) + n(t) \\
&= A_c m(t)\cos(2\pi f_c t) + n_c(t)\cos(2\pi f_c t) - n_s(t)\sin(2\pi f_c t)
\end{aligned}
$$

- Demodulate the received signal by first multiplying r(t) by a locally generated sinusoid $\cos(2\pi f_c t + \phi)$, where ϕ is the phase of the sinusoid.
- Then passing the product signal through an ideal lowpass filter having a bandwidth W.

Effect of Noise on DSB-SC AM

- The multiplication of r(t) with cos(2πfct + ϕ) yields

$$
\begin{aligned}
& r(t)\cos(2\pi f_c t+\phi) \\
&\quad = u(t)\cos(2\pi f_c t+\phi)+n(t)\cos(2\pi f_c t+\phi) \\
&\quad = A_c m(t)\cos(2\pi f_c t)\cos(2\pi f_c t+\phi) \\
&\qquad + n_c(t)\cos(2\pi f_c t)\cos(2\pi f_c t+\phi)-n_s(t)\sin(2\pi f_c t)\cos(2\pi f_c t+\phi) \\
&\quad = \tfrac{1}{2}A_c m(t)\cos(\phi)+\tfrac{1}{2}A_c m(t)\cos(4\pi f_c t+\phi) \\
&\qquad + \tfrac{1}{2}\left[n_c(t)\cos(\phi)+n_s(t)\sin(\phi)\right]+\tfrac{1}{2}\left[n_c(t)\cos(4\pi f_c t+\phi)-n_s(t)\sin(4\pi f_c t+\phi)\right]
\end{aligned}
$$

- The lowpass filter rejects the double frequency components and passes only the lowpass components.

$$
y(t)=\tfrac{1}{2}A_c m(t)\cos(\phi)+\tfrac{1}{2}\left[n_c(t)\cos(\phi)+n_s(t)\sin(\phi)\right]
$$

Effect of Noise on DSB-SC AM

- In Chapter 3, the effect of a phase difference between the received carrier and a locally generated carrier at the receiver is a drop equal to $cos^2(\phi)$ in the received signal power.
- Phase-locked loop (Section 6.4)
  - The effect of a phase-locked loop is to generate phase of the received carrier at the receiver.
  - If a phase-locked loop is employed, then $\phi = 0$ and the demodulator is called a coherent or synchronous demodulator.
- In our analysis in this section, we assume that we are employing a coherent demodulator.
  - With this assumption, we assume that $\phi = 0$

$$y(t) = \tfrac{1}{2}\left[A_c m(t) + n_c(t)\right]$$

Effect of Noise on DSB-SC AM

- Therefore, at the receiver output, the message signal and the noise components are additive and we are able to define a meaningful SNR. The message signal power is given by

$$P_o = \frac{A_c^2}{4} P_M$$

  - power P_M is the content of the message signal

- The noise power is given by

$$P_{n_0} = \frac{1}{4} P_{n_c} = \frac{1}{4} P_n$$

- The power content of n(t) can be found by noting that it is the result of passing $n_w(t)$ through a filter with bandwidth B_c.

Effect of Noise on DSB-SC AM

- Therefore, the power spectral density of n(t) is given by

$$S_n(f) = \begin{cases} \frac{N_0}{2} & |f - f_c| < W \\ 0 & otherwise \end{cases}$$

- The noise power is

$$P_n = \int_{-\infty}^{\infty} S_n(f)df = \frac{N_0}{2} \times 4W = 2WN_0$$

- Now we can find the output SNR as

$$\left(\frac{S}{N}\right)_0 = \frac{P_0}{P_{n_0}} = \frac{\frac{A_c^2}{4} P_M}{\frac{1}{4} 2WN_0} = \frac{A_c^2 P_M}{2WN_0}$$

- In this case, the received signal power, given by Eq. (3.2.2), is

$$P_R = A_c^2 P_M / 2.$$

Effect of Noise on DSB-SC AM

- The output SNR for DSB-SC AM may be expressed as

$$\left(\frac{S}{N}\right)_{0_{DSB}} = \frac{P_R}{N_0 W}$$

  - which is identical to baseband SNR which is given by Equation (6.1.2).

- In DSB-SC AM, the output SNR is the same as the SNR for a baseband system

  $\Rightarrow$ DSB-SC AM does not provide any SNR improvement over a simple baseband communication system

Effect of Noise on SSB AM

- SSB modulated signal :

$$u(t) = A_c m(t)\cos(2\pi f_c t) \mp A_c \hat{m}(t)\sin(2\pi f_c t)$$

- Input to the demodulator

$$\begin{aligned} r(t) &= A_c m(t)\cos(2\pi f_c t) \mp A_c \hat{m}(t)\sin(2\pi f_c t) + n(t) \\ &= A_c m(t)\cos(2\pi f_c t) \mp A_c \hat{m}(t)\sin(2\pi f_c t) + n_c(t)\cos(2\pi f_c t) - n_s(t)\sin(2\pi f_c t) \\ &= \left[A_c m(t) + n_c(t)\right]\cos(2\pi f_c t) + \left[\mp A_c \hat{m}(t) - n_s(t)\right]\sin(2\pi f_c t) \end{aligned}$$

- Assumption : Demodulation with an ideal phase reference.
- Hence, the output of the lowpass filter is the in-phase component (with a coefficient of ½) of the preceding signal.

$$y(t) = \tfrac{1}{2}\left[A_c m(t) + n_c(t)\right]$$

Effect of Noise on SSB AM

- Parallel to our discussion of DSB, we have

$$P_o = \frac{A_c^2}{4} P_M$$

$$P_{n_0} = \frac{1}{4} P_{n_c} = \frac{1}{4} P_n$$

$$P_n = \int_{-\infty}^{\infty} S_n(f)\,df = \frac{N_0}{2} \times 2W = WN_0$$

$$\left(\frac{S}{N}\right)_0 = \frac{P_0}{P_{n_0}} = \frac{A_c^2 P_M}{WN_0}$$

$$P_R = P_U = A_c^2 P_M$$

$$\left(\frac{S}{N}\right)_{0_{SSB}} = \frac{P_R}{N_0 W} = \left(\frac{S}{N}\right)_b$$

- The signal-to-noise ratio in an SSB system is equivalent to that of a DSB system.

Effect of Noise on Conventional AM

- DSB AM signal : $u(t) = A_c[1 + am_n(t)]\cos(2\pi f_c t)$
- Received signal at the input to the demodulator

$$
\begin{aligned}
r(t) &= A_c[1 + am_n(t)]\cos(2\pi f_c t) + n(t) \\
&= A_c[1 + am_n(t)]\cos(2\pi f_c t) + n_c(t)\cos\left(2\pi f_c t\right) - n_s(t)\sin\left(2\pi f_c t\right) \\
&= \left[A_c[1 + am_n(t)] + n_c(t)\right]\cos(2\pi f_c t) - n_s(t)\sin\left(2\pi f_c t\right)
\end{aligned}
$$

  - a is the modulation index
  - $m_n(t)$ is normalized so that its minimum value is -1
  - If a synchronous demodulator is employed, the situation is basically similar to the DSB case, except that we have $1 + am_n(t)$ instead of $m(t)$.
- After mixing and lowpass filtering

$$
y(t) = \tfrac{1}{2}\left[A_c a m_n(t) + n_c(t)\right]
$$

Effect of Noise on Conventional AM

- Received signal power

$$P_R = \frac{A_c^2}{2}\left[1 + a^2 P_{M_n}\right]$$

  - Assumed that the message process is zero mean.

- Now we can derive the output SNR as

$$\left(\frac{S}{N}\right)_{0_{AM}} = \frac{\frac{1}{4}A_c^2 a^2 P_{M_n}}{\frac{1}{4}P_{n_c}} = \frac{A_c^2 a^2 P_{M_n}}{2N_0 W} = \frac{a^2 P_{M_n}}{1 + a^2 P_{M_n}} \frac{\frac{A_c^2}{2}\left[1 + a^2 P_{M_n}\right]}{N_0 W}$$

$$= \frac{a^2 P_{M_n}}{1 + a^2 P_{M_n}} \frac{P_R}{N_0 W} = \frac{a^2 P_{M_n}}{1 + a^2 P_{M_n}} \left(\frac{S}{N}\right)_b = \eta \left(\frac{S}{N}\right)_b$$

  - η denotes the modulation efficiency
  - Since $a^2 P_{M_n} < 1 + a^2 P_{M_n}$, the SNR in conventional AM is always smaller than the SNR in a baseband system.

Effect of Noise on Conventional AM

- In practical applications, the modulation index a is in the range of 0.8-0.9.
- Power content of the normalized message process depends on the message source.
- Speech signals : Large dynamic range, P_M is about 0.1.
  - The overall loss in SNR, when compared to a baseband system, is a factor of 0.075 or equivalent to a loss of 11 dB.
- **The reason for this loss** is that a large part of the transmitter power is used to send the carrier component of the modulated signal and not the desired signal.

- To analyze the envelope-detector performance in the presence of noise, we must use certain approximations.
  - This is a result of the nonlinear structure of an envelope detector, which makes an exact analysis difficult.

Effect of Noise on Conventional AM

- In this case, the demodulator detects the envelope of the received signal and the noise process.
- The input to the envelope detector is

$$r(t) = \left[A_c[1+am_n(t)] + n_c(t)\right]\cos(2\pi f_c t) - n_s(t)\sin\left(2\pi f_c t\right)$$

- Therefore, the envelope of $r(t)$ is given by

$$V_r(t) = \sqrt{\left[A_c[1+am_n(t)] + n_c(t)\right]^2 + n_s^2(t)}$$

- Now we assume that the signal component in $r(t)$ is much stronger than the noise component. Then

$$P\left(n_c(t) << A_c[1+am_n(t)]\right) \approx 1$$

- Therefore, we have a high probability that

$$V_r(t) \approx A_c[1+am_n(t)] + n_c(t)$$

Effect of Noise on Conventional AM

- After removing the DC component, we obtain

$$y(t) = A_c a m_n(t) + n_c(t)$$

  - which is basically the same as y*(t)* for the synchronous demodulation without the ½ coefficient.
  - This coefficient, of course, has no effect on the final SNR.
  - So we conclude that, **under the assumption of high SNR at the receiver input, the performance of synchronous and envelope demodulators is the same**.
- However, if the preceding assumption is not true, that is, if we assume that, at the receiver input, the noise power is much stronger than the signal power, Then

Effect of Noise on Conventional AM

$$V_r(t) = \sqrt{[A_c[1+am_n(t)]+n_c(t)]^2 + n_s^2(t)}$$

$$= \sqrt{A_c^2[1+am_n(t)]^2 + n_c^2(t) + n_s^2(t) + 2A_c n_c(t)[1+am_n(t)]}$$

$$\xrightarrow{a} \sqrt{\left(n_c^2(t)+n_s^2(t)\right)\left[1+\frac{2A_c n_c(t)}{n_c^2(t)+n_s^2(t)}\left(1+am_n(t)\right)\right]}$$

$$\xrightarrow{b} V_n(t)\left[1+\frac{A_c n_c(t)}{V_n^2(t)}\left(1+am_n(t)\right)\right]$$

$$= V_n(t) + \frac{A_c n_c(t)}{V_n(t)}\left(1+am_n(t)\right)$$

- (a) : $A_c^2[1+am_n(t)]^2$ is small compared with the other components
- (b) : $\sqrt{n_c^2(t)+n_s^2(t)} = V_n(t)$;the envelope of the noise process
- Use the approximation

  $\sqrt{1+\varepsilon} \approx 1+\frac{\varepsilon}{2}$, *for small* ε, where $\varepsilon = \frac{2A_c n_c(t)}{n_c^2(t)+n_s^2(t)}\left(1+am_n(t)\right)$

Effect of Noise on Conventional AM

- Then

$$V_r(t) = V_n(t) + \frac{A_c n_c(t)}{V_n(t)}\left(1 + a m_n(t)\right)$$

  - We observe that, at the demodulator output, the signal and the noise components are **no longer additive**.
  - In fact, *the signal component is multiplied by noise* and is no longer distinguishable.
  - In this case, no meaningful SNR can be defined.
  - We say that this system is *operating below the threshold.*
  - The subject of threshold and its effect on the performance of a communication system will be covered in more detail when we discuss the noise performance in angle modulation.

Department of ECE

Sub: Analog Communications

Unit - 5

PUSLE MODULATION TECHNIQUES

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Content:

- Pulse Amplitude Modulation
- Pulse Width and Position Modulation
- Demodulation of PWM & PPM

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

- **The circuit of Figure 11-3 is used to illustrate pulse amplitude modulation (PAM). The FET is the switch used as a sampling gate.**
- **When the FET is on, the analog voltage is shorted to ground; when off, the FET is essentially open, so that the analog signal sample appears at the output.**
- **Op-amp 1 is a noninverting amplifier that isolates the analog input channel from the switching function.**

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

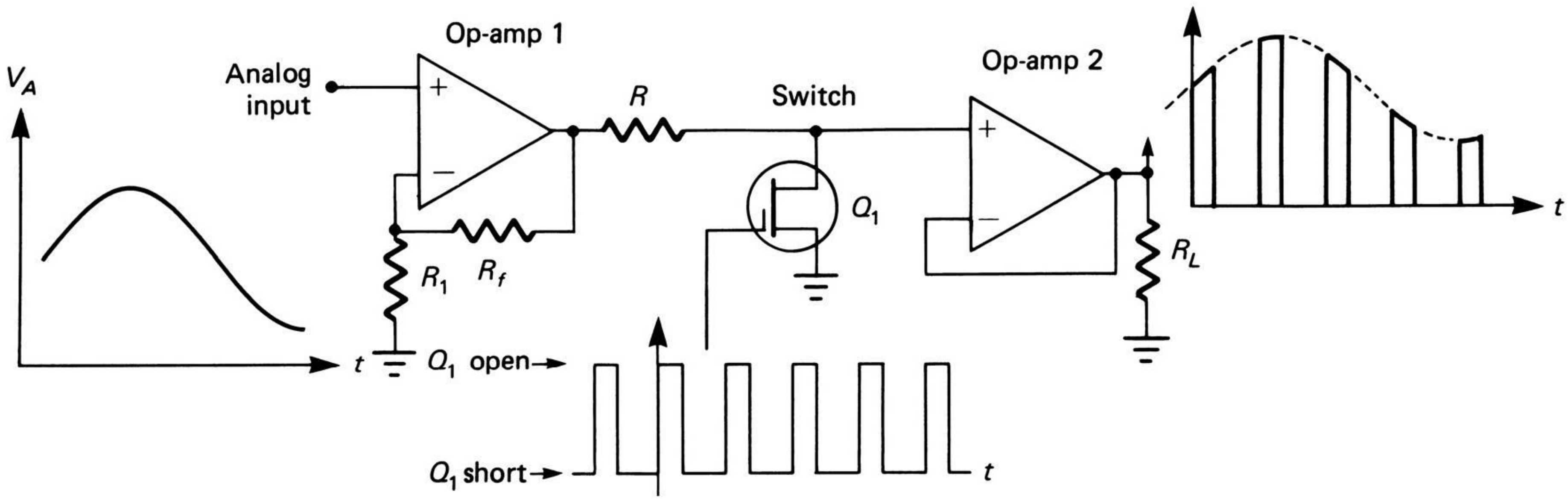


圖 11-3　脈衝幅度調變器，自然取樣。

Figure 11-3. Pulse amplitude modulator, natural sampling.

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

- **Op-amp 2 is a high input-impedance voltage follower capable of driving low-impedance loads (high "fanout").**
- **The resistor *R* is used to limit the output current of op-amp 1 when the FET is "on" and provides a voltage division with r_d of the FET. (r_d, the drain-to-source resistance, is low but not zero)**

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

- **The most common technique for sampling voice in PCM systems is to a sample-and-hold circuit.**
- **As seen in Figure 11-4, the instantaneous amplitude of the analog (voice) signal is held as a constant charge on a capacitor for the duration of the sampling period T_s.**
- **This technique is useful for holding the sample constant while other processing is taking place, but it alters the frequency spectrum and introduces an error, called aperture error, resulting in an inability to recover exactly the original analog signal.**

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

- **The amount of error depends on how mach the analog changes during the holding time, called aperture time.**
- **To estimate the maximum voltage error possible, determine the maximum slope of the analog signal and multiply it by the aperture time ΔT in Figure 11-4.**

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

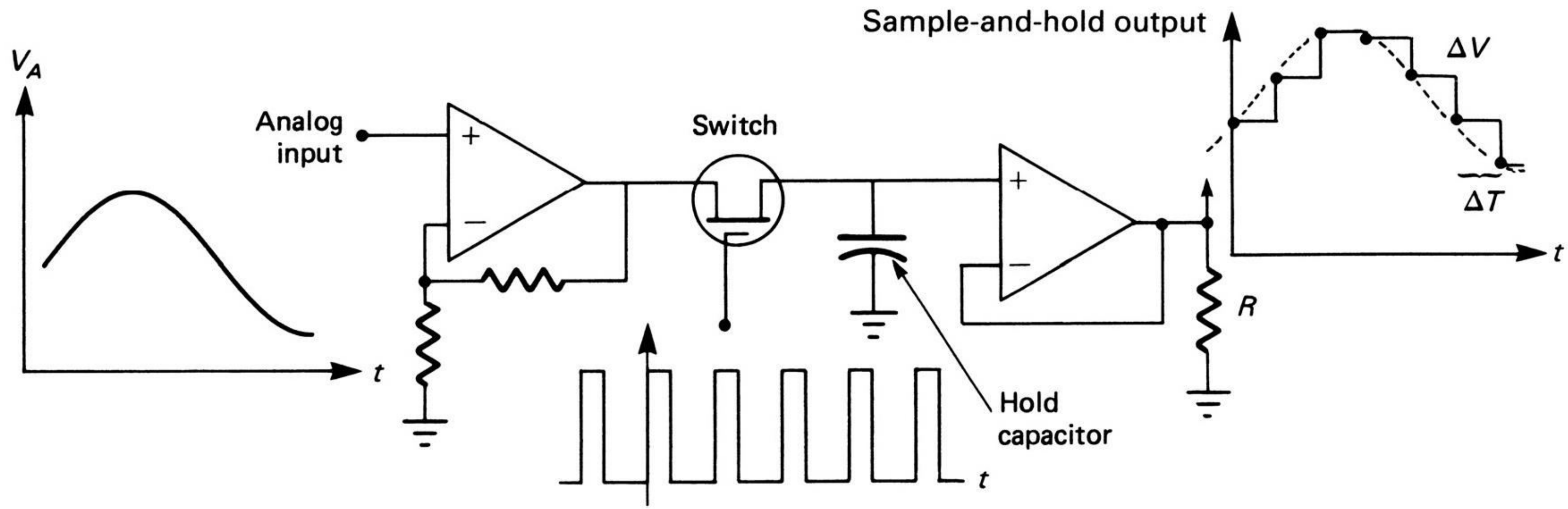


圖 11-4　樣本－和－持保電路和平頂取樣。

Figure 11-4. Sample-and-hold circuit and flat-top sampling.

Pulse Amplitude Modulation – Natural and Flat-Top Sampling

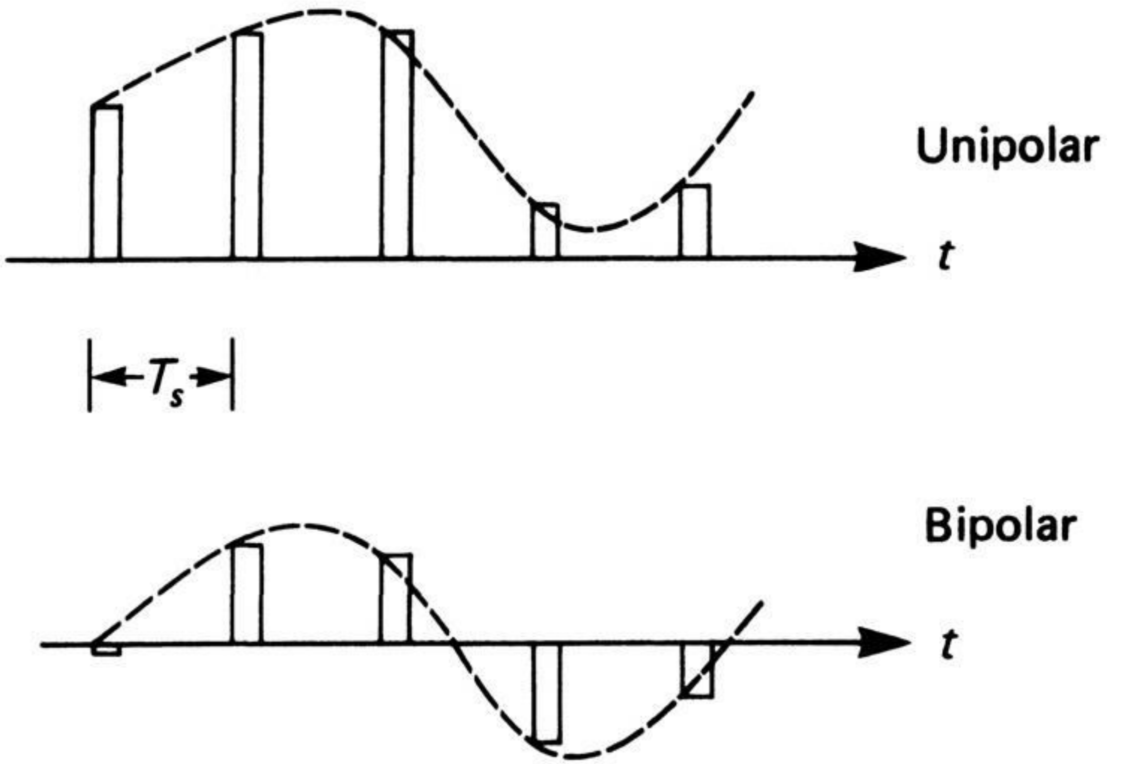
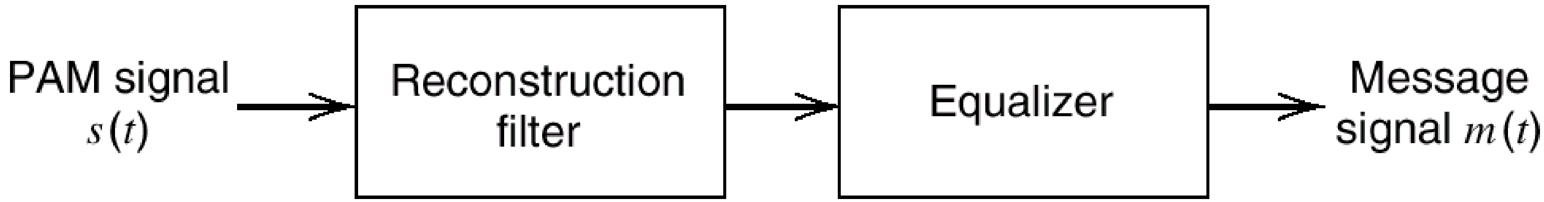


圖 11-5 平頂 PAM 訊號。

Figure 11-5. Flat-top PAM signals.

Recovering the original message signal m(t) from PAM signal



Where the filter bandwidth is W

The filter output is $f_s M(f)H(f)$. Note that the Fourier transform of $h(t)$ is given by

$$H(f) = T\,\text{sinc}(fT)\exp(-j\pi f T) \qquad (3.19)$$

amplitude distortion — delay $= T/2$

$\Rightarrow$ aparture effect

Let the equalizer response is

$$\frac{1}{H(f)} = \frac{1}{T\,\text{sinc}(fT)} = \frac{\pi f}{\sin(\pi f T)} \qquad (3.20)$$

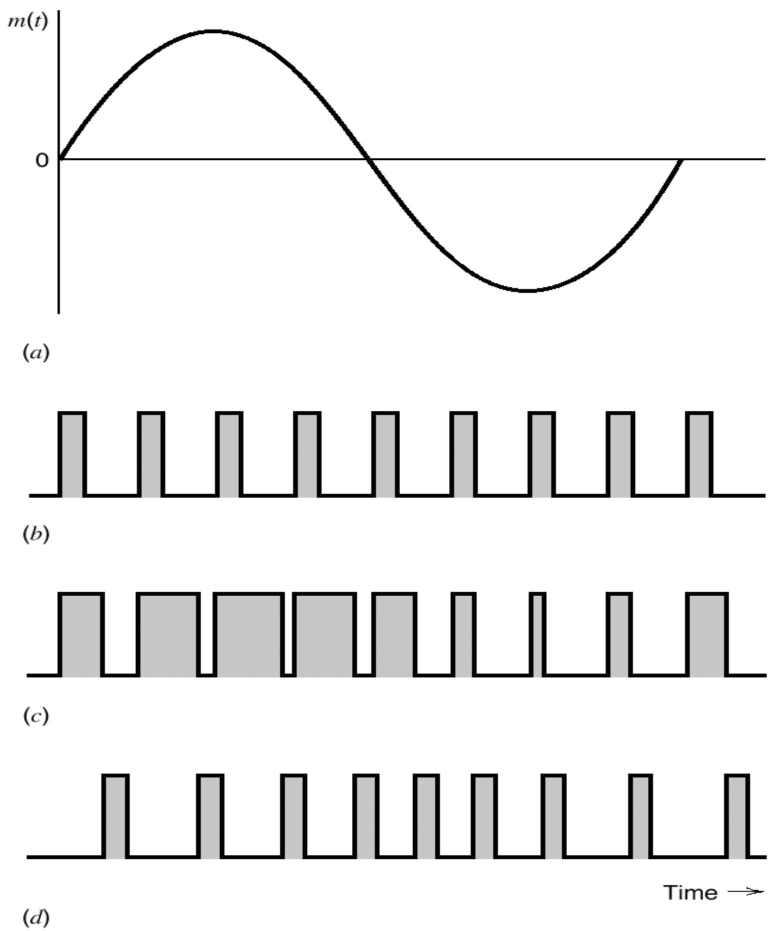
Ideally the original signal $m(t)$ can be recovered completely.

3.4 Other Forms of Pulse Modulation

a. Pulse-duration modulation (PDM)

b. Pulse-position modulation (PPM)

PPM has a similar noise performance as FM.



Pulse Width and Pulse Position Modulation

- In pulse width modulation (PWM), the width of each pulse is made directly proportional to the amplitude of the information signal.
- In pulse position modulation, constant-width pulses are used, and the position or time of occurrence of each pulse from some reference time is made directly proportional to the amplitude of the information signal.
- PWM and PPM are compared and contrasted to PAM in Figure 11-11.

Pulse Width and Pulse Position Modulation

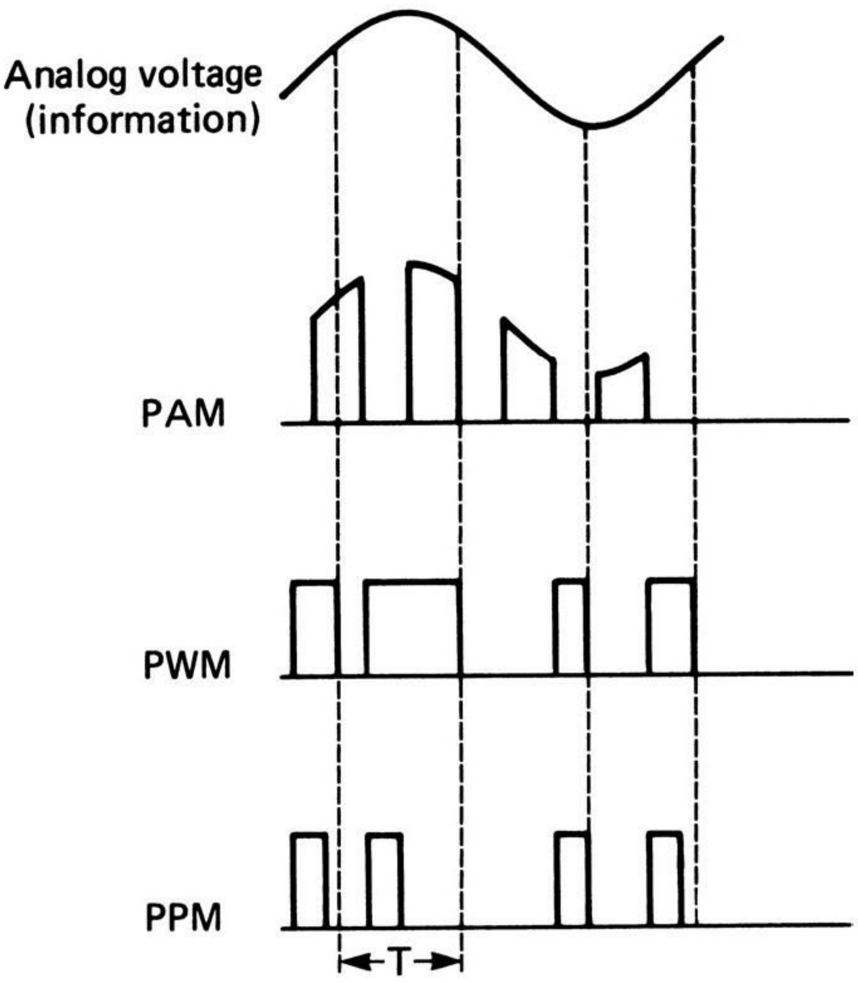


圖 11-11　類比脈衝調變訊號。

Figure 11-11. Analog/pulse modulation signals.

Pulse Width and Pulse Position Modulation

- Figure 11-12 shows a PWM modulator. This circuit is simply a high-gain comparator that is switched on and off by the sawtooth waveform derived from a very stable-frequency oscillator.
- Notice that the output will go to $+V_{cc}$ the instant the analog signal exceeds the sawtooth voltage.
- The output will go to $-V_{cc}$ the instant the analog signal is less than the sawtooth voltage. With this circuit the average value of both inputs should be nearly the same.
- This is easily achieved with equal value resistors to ground. Also the $+V$ and $-V$ values should not exceed V_{cc}.

Pulse Width and Pulse Position Modulation

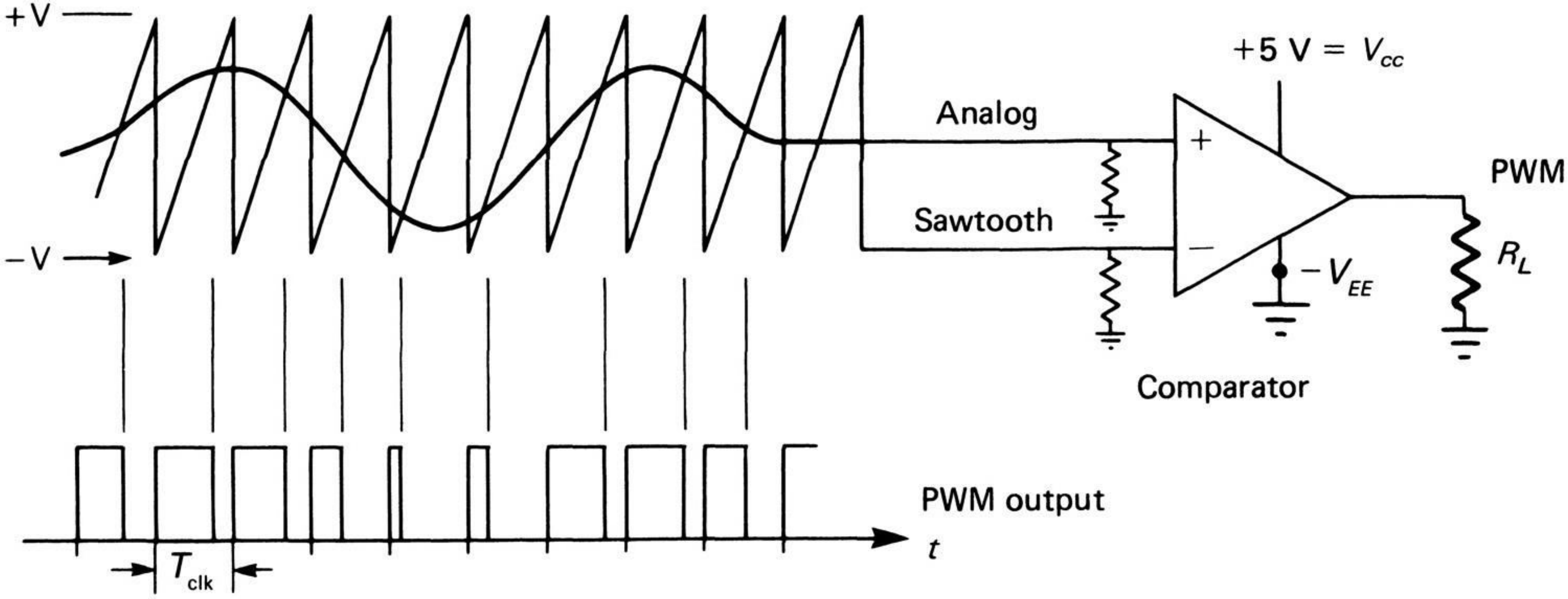


圖 11-12 脈衝寬度調變器。

Figure 11-12. Pulse width modulator.

Pulse Width and Pulse Position Modulation

- **A 710-type IC comparator can be used for positive-only output pulses that are also TTL compatible. PWM can also be produced by modulation of various voltage-controllable multivibrators.**
- **One example is the popular 555 timer IC. Other (pulse output) VCOs, like the 566 and that of the 565 phase-locked loop IC, will produce PWM.**
- **This points out the similarity of PWM to continuous analog FM. Indeed, PWM has the advantages of FM---constant amplitude and good noise immunity---and also its disadvantage---large bandwidth.**

Demodulation

- Since the width of each pulse in the PWM signal shown in Figure 11-13 is directly proportional to the amplitude of the modulating voltage.
- The signal can be differentiated as shown in Figure 11-13 (to PPM in part a), then the positive pulses are used to start a ramp, and the negative clock pulses stop and reset the ramp.
- This produces frequency-to-amplitude conversion (or equivalently, pulse width-to-amplitude conversion).
- The variable-amplitude ramp pulses are then time-averaged (integrated) to recover the analog signal.

Pulse Width and Pulse Position Modulation

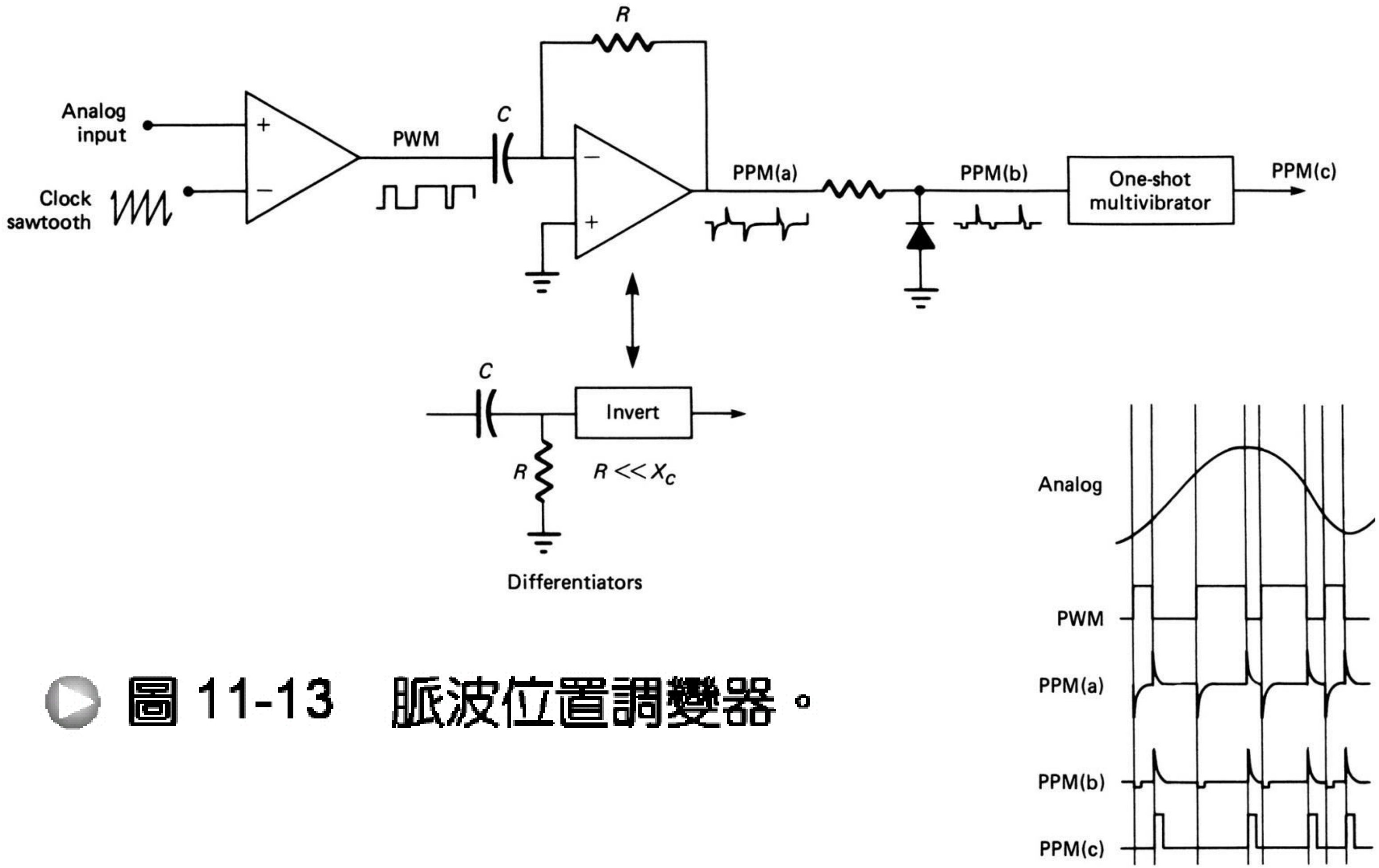


圖 11-13　脈波位置調變器。

Figure 11-13. Pulse position modulator.

Demodulation

- As illustrated in Figure 11-14, a narrow clock pulse sets an RS flip-flop output high, and the next PPM pulses resets the output to zero.
- The resulting signal, PWM, has an average voltage proportional to the time difference between the PPM pulses and the reference clock pulses.
- Time-averaging (integration) of the output produces the analog variations.
- PPM has the same disadvantage as continuous analog phase modulation: a coherent clock reference signal is necessary for demodulation.
- The reference pulses can be transmitted along with the PPM signal.

Demodulation

- This is achieved by full-wave rectifying the PPM pulses of Figure 11-13a, which has the effect of reversing the polarity of the negative (clock-rate) pulses.
- Then an edge-triggered flipflop (J-K or D-type) can be used to accomplish the same function as the RS flip-flop of Figure 11-14, using the clock input.
- The penalty is: more pulses/second will require greater bandwidth, and the pulse width limit the pulse deviations for a given pulse period.

Demodulation

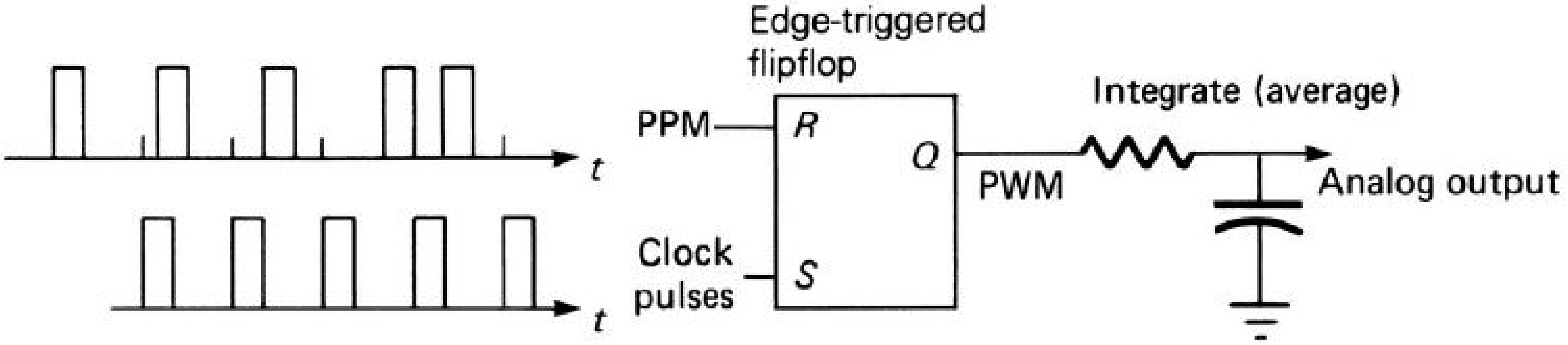


圖 11-14　PPM 解調器。

Figure 11-14. PPM demodulator.