Phase Modulation (PM)
- In PM, excess phase $\Delta \phi(t)$ of the carrier is varied linearly with the message signal $s(t)$. That is,
  \[ \theta(t) = k_p s(t) = k_p \max[|s(t)|] s(t) = \Delta \phi_{\text{max}} s(t) \]
  where
  \[ k_p = \text{Phase sensitivity of the phase modulator (radians/volt)} \]
  \[ s_n(t) = \max[|s(t)|] s(t) \]
  \[ \Delta \phi_{\text{max}} = \max[|\theta(t)|] = k_p \max[|s(t)|] \]
- Maximum phase shift or deviation produced by the message signal $s(t)$
  \[ x_{\text{pm}}(t) = A_0 \cos[2\pi f + 2\Delta \phi_{\text{max}} s(t)] \]

Frequency Modulation (FM)
- In FM, the instantaneous frequency $f_i(t)$ of the carrier is varied linearly with the message signal $s(t)$. That is,
  \[ f_i(t) = f_c + k_f \max[|s(t)|] s(t) = f_c + \Delta f_{\text{max}} s(t) \]
  where
  \[ k_f = \text{Frequency sensitivity of the FM modulator (Hz/volt)} \]
  \[ \Delta f_{\text{max}} = \text{Maximum frequency deviation from } f_c \]
- Instantaneous phase of the FM signal
  \[ \phi(t) = 2\pi f_c t + 2\pi \Delta f_{\text{max}} \int_{-\infty}^{t} s_n(\alpha) d\alpha \]
- FM signal
  \[ x_{\text{fm}}(t) = A_0 \cos[2\pi f_c t + 2\pi \Delta f_{\text{max}} \int_{-\infty}^{t} s_n(\alpha) d\alpha] \]

Relationship between FM and PM
- An FM waveform corresponding to the message signal $s_n(t)$ is also a PM waveform corresponding to the signal $s_n(t)$.
- Similarly, a PM waveform corresponding to message signal $s_n(t)$ is also an FM waveform corresponding to signal $s_n(t)$.

FM and PM Signals: Square Wave Modulating Signal
- Angle Modulation
  - An angle-modulated signal can be expressed as
    \[ x(t) = A_0 \cos[2\pi f_c t + \theta(t)] \]
  - Information-bearing signal embedded in the instantaneous phase/frequency of the carrier
    \[ \text{Excess phase} \]
    \[ \phi(t) = 2\pi f_c t + \theta(t) \]
    \[ \text{Instantaneous frequency of } x(t): \quad f_i(t) = f_c + \frac{1}{2\pi} \frac{d\phi}{dt} \]
- Modulated waveform does not resemble message waveform
- Amplitude is constant $\Rightarrow$ we can use more efficient nonlinear amplifiers

- Frequency Modulation (FM)
  - An angle-modulated signal can be expressed as
    \[ x(t) = A_0 \cos[2\pi f_c t + \theta(t)] \]
  - Information-bearing signal embedded in the instantaneous phase/frequency of the carrier
    \[ \text{Excess phase} \]
    \[ \phi(t) = 2\pi f_c t + \theta(t) \]
    \[ \text{Instantaneous frequency of } x(t): \quad f_i(t) = f_c + \frac{1}{2\pi} \frac{d\phi}{dt} \]
- Modulated waveform does not resemble message waveform
- Amplitude is constant $\Rightarrow$ we can use more efficient nonlinear amplifiers
**FM and PM Signals: Sinusoidal Modulation**

- Let \( s(t) = A_c \cos(2\pi f_c t) \)
- **Instantaneous Phase**
  \[
  \phi(t) = \begin{cases}
    2\pi f_c t + \frac{\Delta f_m}{f_m} \sin(2\pi f_c t) & : \text{FM} \\
    2\pi f_c t + \Delta f_m \cos(2\pi f_c t) & : \text{PM}
  \end{cases}
  \]
- **Modulated signal**
  \[
  x_{FM}(t) = A_c \cos \left(2\pi f_c t + \beta \sin(2\pi f_c t) \right)
  \]
  \[
  x_{PM}(t) = A_c \cos \left(2\pi f_c t + \Delta f_m \cos(2\pi f_c t) \right)
  \]
- **Modulation indices**
  \[
  \text{FM: } \beta = \frac{\Delta f_m}{f_m}, \quad \Delta f_m = k_f A_m
  \]
  \[
  \text{PM: } \beta = \Delta f_m, \quad \Delta f_m = k_f A_m
  \]

**Power in Angle-Modulated Signal**

- The normalized power content of an angle-modulated signal is given by
  \[
  P_t = \lim_{T \to \infty} A^2 \int_{-T/2}^{T/2} \cos^2(2\pi f_c t + \beta(\theta(t)))dt
  \]
  \[
  = A^2 \lim_{T \to \infty} \frac{1}{T} \int_{-T/2}^{T/2} \cos(4\pi f_c t + 2\beta(\theta(t)))dt
  \]
- For \( f_c \) large
  \[
  P_t = \frac{A^2}{2}
  \]
- The average power of an angle-modulated signal is constant independent of the message signal
- Not a surprising result since an angle-modulated signal has constant amplitude

**Spectrum of Sine Wave Modulated FM Signal**

- Consider the sine wave modulated FM signal
  \[
  x_{FM}(t) = A_c \cos \left[2\pi f_c t + \beta \sin(2\pi f_c t) \right] = A_c \Re \left( e^{j2\pi f_c t} e^{j\beta \sin(2\pi f_c t)} \right)
  \]
- The complex envelope \( A_c e^{j(\beta \sin(2\pi f_c t))} \) is periodic with period \( 1/f_m \) and therefore can be expanded in a complex Fourier series:
  \[
  A_c e^{j(\beta \sin(2\pi f_c t))} = A_c \sum_{n=-\infty}^{\infty} C_n e^{j2\pi f_n t}
  \]
- Bessel function of 1st kind and order \( n \)
- Therefore
  \[
  x_{FM}(t) = \Re \left\{ A_c \sum_{n=-\infty}^{\infty} J_n(\beta) e^{j2\pi f_n t} \right\} = A_c \sum_{n=-\infty}^{\infty} J_n(\beta) \cos \left[2\pi (f_n + \beta f_c) t \right]
  \]
- The spectrum of the FM signal not only contains the carrier frequency term \( (n = 0) \) but also components \( (n \neq 0) \) at multiples of the signal frequency on both sides of the carrier
- These components at \( f_c \pm n f_m \) \( n = 1, 2, \ldots \) are called **sidebands** of the FM signal
- The magnitudes of the sidebands in the FM signal become negligible for sufficiently large \( n \)
- The number of sidebands that account for at least 98% power in the FM signal is given by integer part of \( (\beta + 1) \) as illustrated in Table 5.2
- The **98% power bandwidth** of the FM signal is
  \[
  B_r = 2(\beta + 1)f_m = 2(\Delta f_m + f_m)
  \]
Table 5.2 Values of the Bessel Functions \( J_n(\beta) \)

<table>
<thead>
<tr>
<th>( n )</th>
<th>( \beta = 0.1 )</th>
<th>( \beta = 0.2 )</th>
<th>( \beta = 0.5 )</th>
<th>( \beta = 1 )</th>
<th>( \beta = 2 )</th>
<th>( \beta = 5 )</th>
<th>( \beta = 8 )</th>
<th>( \beta = 10 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.058</td>
<td>0.086</td>
<td>0.338</td>
<td>0.384</td>
<td>0.576</td>
<td>0.604</td>
<td>0.712</td>
<td>0.723</td>
</tr>
<tr>
<td>1</td>
<td>0.099</td>
<td>0.124</td>
<td>0.146</td>
<td>0.152</td>
<td>0.175</td>
<td>0.183</td>
<td>0.191</td>
<td>0.194</td>
</tr>
<tr>
<td>2</td>
<td>0.061</td>
<td>0.085</td>
<td>0.115</td>
<td>0.126</td>
<td>0.146</td>
<td>0.154</td>
<td>0.161</td>
<td>0.164</td>
</tr>
<tr>
<td>3</td>
<td>0.029</td>
<td>0.052</td>
<td>0.080</td>
<td>0.093</td>
<td>0.107</td>
<td>0.114</td>
<td>0.119</td>
<td>0.121</td>
</tr>
<tr>
<td>4</td>
<td>0.001</td>
<td>0.014</td>
<td>0.021</td>
<td>0.026</td>
<td>0.030</td>
<td>0.034</td>
<td>0.038</td>
<td>0.039</td>
</tr>
<tr>
<td>5</td>
<td>0.000</td>
<td>0.005</td>
<td>0.008</td>
<td>0.010</td>
<td>0.011</td>
<td>0.012</td>
<td>0.013</td>
<td>0.014</td>
</tr>
<tr>
<td>6</td>
<td>0.001</td>
<td>0.007</td>
<td>0.009</td>
<td>0.010</td>
<td>0.010</td>
<td>0.010</td>
<td>0.010</td>
<td>0.010</td>
</tr>
<tr>
<td>7</td>
<td>0.001</td>
<td>0.004</td>
<td>0.004</td>
<td>0.004</td>
<td>0.003</td>
<td>0.002</td>
<td>0.002</td>
<td>0.002</td>
</tr>
<tr>
<td>8</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
<td>0.000</td>
</tr>
<tr>
<td>9</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>10</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Demodulation of FM Signals

- **Frequency discrimination** – FM signal converted into AM signal by a differentiator. Use an envelope detector to recover the message signal
  - Noncoherent demodulation method
- **Phase discrimination** – FM signal converted into a PM signal. Use a phase detector to recover the message signal.
  - In practice, implemented using a quadrature detector.
  - Noncoherent demodulation method
- **Phase lock loop (PLL) detector**. A PLL detector uses a voltage-controlled oscillator (VCO) and feedback to extract the message signal.
  - Coherent demodulation method

Bandpass Limiter

- Although amplitude of an FM carrier is constant, the signal entering the FM demodulator may have amplitude variations due to addition of the channel noise
- All FM discriminators are, therefore, preceded by a **BP limiter** to ensure that the discriminator input signal is constant in amplitude

FM Magnitude spectra for different values of \( \beta \)

- For a narrowband FM system, \( \beta < 1 \)
  \[ x_{FM}(t) = A_c \cos[2\pi f_c t + \beta \sin(2\pi f_m t)] \]
  \[ = A_c \cos(2\pi f_c t) \frac{\beta}{2} \cos[2\pi (f_c + f_m) t] - \frac{\beta}{2} \cos[2\pi (f_c - f_m) t] \]
- Similar to AM signal, the bandwidth is (both, PM & FM)
  \[ B_f = 2B \]

Narrowband FM

- For a narrowband FM system, \( \beta < < 1 \)
  \[ x_{FM}(t) = A_c \cos[2\pi f_c t + \beta \sin(2\pi f_m t)] \]
  \[ = A_c \cos(2\pi f_c t) \frac{\beta}{2} \cos[2\pi (f_c + f_m) t] - \frac{\beta}{2} \cos[2\pi (f_c - f_m) t] \]

Bandwidth of FM Signal for Arbitrary Message Signal

- For an arbitrary message signal \( s(t) \) having the bandwidth \( B \), we define the **deviation ratio**
  \[ D = \frac{\text{Peak frequency deviation}}{\text{Bandwidth of the message signal}} = \frac{\Delta f_{\text{max}}}{B} \]
- \( D \) plays same role in determining the FM signal bandwidth for an arbitrary modulating signal as the modulation index \( \beta \) plays for the sinusoidal message signal
- The bandwidth of FM signal for arbitrary modulating signal is
  \[ B_f = 2(D + 1)B = 2(\Delta f_{\text{max}} + B) \]
- Similarly, the bandwidth of a PM signal is
  \[ B_p = 2(\Delta f_{\text{max}} + 1)B \]
Frequency Discriminator

- Ideal Differentiator
  \[ H_{fd}(f) = K_{fd}2\pi f \]
  \[ K_{fd} \frac{d}{dt} = \frac{d}{dt} \]

- If the FM signal \( x(t) = A_0 \cos[2\pi f_c t + \theta(t)] \) is input to the differentiator circuit, the output is given by
  \[ v(t) = K_{fd} \frac{dx(t)}{dt} = K_{fd}2\pi f_c \cos[2\pi f_c t + \theta(t)] \]

- In a practice, \( 2\pi f_c + \frac{d\theta(t)}{dt} > 0 \). Hence
  Envelope detector output: \[ y(t) = K_{fd} \left[ 2\pi f_c + \frac{d\theta(t)}{dt} \right] \]

Frequency Discriminator (contd)

- For an FM signal, \( \theta(t) = 2\pi f_c \int x(t) \, dt \)
  Envelope detector output: \[ y(t) = K_{fd} 2\pi s_c(t) \]

- After DC block, the FM discriminator output is
  \[ y_{fd}(t) = K_{fd} \frac{d\theta(t)}{dt} = K_{fd} 2\pi f_c s_c(t) \]

- The discriminator output is proportional to the message signal \( s(t) \)

Slope Detector

- In practice, the ideal differentiator is approximated by a circuit that has a linear amplitude response over the bandwidth of the FM signal as described by
  \[
  |H_{slope,k}(f)| = \begin{cases} 
  2\pi K_{fd} & \text{if } f - \frac{B_c}{2} \leq |f| \leq f + \frac{B_c}{2} \\
  0 & \text{otherwise}
  \end{cases}
  \]

- The magnitude of the frequency response function is linear over the narrow band

- To extend the frequency range, a balanced slope detector is used with two tuned circuits

FM discriminator: Balanced slope detector

Phase-shift Discriminator: Quadrature Detector

- A quadrature detector converts instantaneous frequency deviation in an FM signal to phase shift and then detects the change of phase
  \[ \Delta \phi = \frac{\pi}{2} K_{fd} [f_c(t) - f_c] = \frac{\pi}{2} K_{fd} f_c s_c(t) \]
  \[ \Delta \phi = \frac{\pi}{2} K_{fd} [f_c(t) - f_c] + K_{fd} f_c s_c(t) \]

- The transfer characteristic of the phase-shift network is described by
  \[ \Delta \phi = \frac{\pi}{2} K_{fd} [f_c(t) - f_c] + K_{fd} f_c s_c(t) \]

- Note that the phase shift is linearly proportional to the instantaneous frequency deviation from \( f_c \)
Quadrature Detector (contd)
- Output of the phase-shift network to input FM signal
  \[ x(t) = K_A \cos \left( 2\pi f + 2\pi f_{\text{mod}} \int_{-\infty}^{t} x(t') \, dt' + K_A f_{\text{mod}} x(t) \right) \]
- The phase detector is implemented by cascade of a mixer and LP filter. Its output is given by
  \[ v(t) = K_v K_A \cos \left( 2\pi f + 2\pi f_{\text{mod}} \int_{-\infty}^{t} x(t') \, dt' + K_A f_{\text{mod}} x(t) \right) \]
  - For sufficiently small \( K \), the output becomes
  \[ y(t) = \frac{K_v K_A}{2} \cos \left( 2\pi f + 2\pi f_{\text{mod}} \int_{-\infty}^{t} x(t') \, dt' + K_A f_{\text{mod}} x(t) \right) \]

Conventional AM versus FM/PM
- AM is linear, FM or PM is highly nonlinear
- Conventional AM is simple (envelope detector) but no noise/interference immunity (low quality).
- AM bandwidth is twice or the same as the modulating signal (no bandwidth expansion).
- Power efficiency—low for conventional AM.
- DSB-SC & SSB—good power efficiency, but complex circuitry.
- FM/PM provide many advantages
  - Noise immunity over AM, at the cost of larger bandwidth
  - Good quality
  - More complex circuitry. However, ICs allow for cost effective implementation

Phase-Locked Loop (PLL): Introduction
- PLL is one of the most versatile circuit blocks used in both communication and instrumentation systems.
- PLLs are widely used in cell phones, televisions, radios, pagers, computers, and storage devices
- A PLL contains three basic components as shown below

PLL: Introduction (contd)
- Under the phase-locked condition, a PLL satisfies the following properties:
  \[ \theta_{\text{ref}} = \theta_{\text{out}} + \text{constant} \]
  - \( \theta_{\text{ref}} \) = Frequency of the input reference signal (rad/s)
  - \( \theta_{\text{out}} \) = Frequency of the oscillator output (rad/s)
  - \( \theta_{\text{ref}} \) = Excess phase of the input reference signal
  - \( \theta_{\text{out}} \) = Excess phase of the oscillator output
- PLL Types

Operation of the PLL
- PD compares phases of the input signal \( x(t) \) and the VCO output \( v_{\text{out}}(t) \) to produce an error signal \( v_{\text{err}}(t) \)
- This error signal is then filtered – remove noise and unwanted frequency components to produce the control voltage \( v_{\text{cont}}(t) \)
- The frequency of the VCO is varied in accordance with \( v_{\text{cont}}(t) \) until the phase lock is achieved
- Once the PLL has acquired the lock, the VCO will track the input reference signal frequency over some range, provided that the input frequency changes slowly
- The PLL output can be taken from either \( v_{\text{cont}}(t) \), the filtered (baseband) VCO control voltage, or the output of the VCO, depending on the application
  - The baseband output tracks the phase variation at the input
  - The VCO output can be used as a LO or clock signal
Analog Phase-Locked Loop (APLL)

- PLL input
  \[ x(t) = A \cos(\omega t + \theta_e(t)) \]

- VCO output
  \[ v_{oc}(t) = B \sin(\omega t + \theta_e(t)) \]

- For APLL, PD is multiplier followed by LP filter. Its output is
  \[ v_p(t) = \frac{1}{2} K_B K_v \sin[\theta_e(t) - \theta_e(t)] \]

It is convenient to express the PD output as

\[ \psi(t) = v_p(t) = K_{PD} \sin(\theta_e(t)) \]

PD gain constant (volts/radian). Slope of the PD transfer characteristic at \( \theta_e = 0 \).

VCO

- The VCO oscillates at an instantaneous frequency that is linearly proportional to the control voltage \( v_{cont}(t) \). That is,
  \[ \omega(t) = \omega_0 K_{VCO} v_{cont}(t) \]

- \( \omega_0 \) is the free-running frequency of the VCO (\( v_{cont}(t) = 0 \))

- \( K_{VCO} \) is the VCO gain constant in radians/sec-volt

- The excess phase of the VCO output is
  \[ \Theta_e(t) = K_{VCO} \int v_{cont}(t) dt \]

- In \( s \)-domain
  \[ \Theta_e(s) = \frac{V_{cont}(s)K_{VCO}}{s} \]

APLL in Tracking Mode – Linear Model

- Phase error \( \theta_e \) small \( \Rightarrow \sin \theta_e = \theta_e \). The PD output is
  \[ v_{p}(t) = K_{PD} \theta_e(t) \]

- Allows to develop a linearized model of the PLL

Open-loop transfer function of the PLL

\[ G(s) = \frac{K_{PD} F(s)K_{VCO}}{s} \]

Loop gain \( K = K_{PD} K_{VCO} F(0) \)

Closed-loop transfer function of the PLL

\[ H(s) = \frac{\Theta_e(s)}{\Theta_e(s)} = \frac{G(s)}{1 + G(s)} = \frac{K F(s)}{s + K F(s)} \]

APLL – Linear Model (contd)

- Phase error transfer function
  \[ H_e(s) = \frac{\Theta_e(s)}{\Theta_e(s)} - \frac{\Theta_e(s) - \Theta_{in}(s)}{\Theta_e(s)} = 1 - H(s) = \frac{s}{s + K F(s)} \]

We can now write the following relationship between the phase error and the input excess phase

\[ \Theta_e(s) = \Theta_{in}(s) H_e(s) = \Theta_{in}(s) \frac{s}{s + K F(s)} \]

- Steady-state behavior of the PLL. Applying the final value theorem of the Laplace transform, the steady-state phase error \( \hat{\theta}(s) \) is given by
  \[ \hat{\theta}(s) = \lim_{s \to 0^+} \Theta_e(s) = \lim_{s \to 0^+} \frac{s}{s + K F(s)} = \frac{s \Theta_{in}(s)}{s + K} \]

First-Order PLL

- A loop filter with no poles, that is \( F(s) = 1 \), produces a first-order PLL

Closed-loop transfer function

\[ H(s) = \frac{K}{s + K} \]

For loop stability, \( K \geq 0 \). The 3-dB bandwidth of the loop is \( K \)

- Steady-state error
  - Phase step \( \Delta \theta \), Steady-state phase error \( \hat{\theta}(s) = \lim_{s \to 0^+} \frac{s \Delta \theta}{s + K} = 0 \)

- Frequency step \( \Delta \omega \),
  \[ \text{Steady-state phase error} \ \hat{\theta}(s) = \lim_{s \to 0^+} \frac{\Delta \omega}{s + K} = \frac{\Delta \omega}{K} \]

\( \hat{\theta} \) is smaller if \( K \) large. But more loop noise. Can’t achieve both.
Second-order PLL
- The use of a loop filter with a single pole produces a second-order PLL.
- Three loop filters considered

Transfer Function of Second-order PLL
- A second-order PLL is characterized by the following parameters:
  - $K$: Open-loop gain
  - $\omega_n$: Natural frequency
  - $\zeta$: Damping factor

Frequency Response of a Second-order PLL
- We observe from Figure that a second-order PLL has a low-pass filter characteristic with flat response up to $\omega_n$ (with some peaking determined by the value of damping factor $\zeta$).
- This implies that the PLL can track phase and frequency variations of the reference input signal up to frequency $\omega_n$.
- The damping factor $\zeta$ has significant influence in determining the transient behavior of the PLL.
  - For $\zeta = 1$, the PLL is critically damped.
  - For $\zeta < 1$, the transient response becomes oscillatory, with overshoot becoming larger as $\zeta$ is made smaller. The choice of $\zeta = 0.707$, provides a good compromise between the flatness and the settling behavior for the loop.
  - For $\zeta > 1$, the PLL is over damped. Used in clock recovery circuits.

Steady-state Phase Errors for Various Loop filters
- Table summarizes steady-state phase errors achievable for second-order PLLs with various loop filters.
- We observe from the table that to track the frequency offset with zero phase error, an active PI loop filter is required.

Acquisition Process in the PLL
- If the PLL is initially unlocked, the phase error, $\theta_e$, can take on arbitrarily large values and as a result, the linear model is no longer valid.
- The process of acquiring phase lock from the unlocked state is called the acquisition process.
- Figure displays the VCO output frequency as a function of the PLL input frequency.
Key Operating Frequency Ranges of the PLL

- **Pull-in (or capture) range** ($\Delta f_{pl}$) – input frequency range over which a PLL acquires lock
- **Hold-in range** ($\Delta f_{hi}$) – input frequency range over which a PLL remains locked once it acquires the phase lock
- **Lock-in range** ($\Delta f_{li}$) – input frequency range within which a PLL locks within one single-beat note between the reference input and the VCO output frequencies

PLL as FM Demodulator

- Let the PLL input signal is FM carrier $x_{in}(t) = A \cos(\omega_0 t + \theta_0(t))$

- Taking FT of both sides

$$\Theta_{\omega}(j\omega) = \frac{2\pi \alpha}{\omega_0} \int s(\alpha) d\alpha$$

- The relationship between the VCO control voltage and its output in the frequency domain is given by

$$\Theta_{\omega}(j\omega) = \frac{K_{VCO}}{j\omega} V_{\text{in}}$$

PLL as FM Demodulator (contd)

- That is, 

$$V_{\text{out}}(j\omega) = \frac{\omega_0}{K_{VCO}} H(j\omega) \Theta_{\omega}(j\omega)$$

- For first-order PLL

$$V_{\text{out}}(j\omega) = \frac{2\pi \alpha}{\omega_0 + j\omega + K_{VCO}} S(j\omega)$$

- If 3-dB bandwidth of the loop $\omega_{3dB} = \frac{K_{VCO} \omega_0}{2\pi}$, then

$$V_{\text{out}}(j\omega) = \frac{2\pi \alpha}{\omega_0 + j\omega} S(j\omega)$$

PLL acts as an FM demodulator. The detected output is $\omega_0(t)$. 

FM Radio Broadcasting

- Signal bandwidth =15 kHz
- Frequency band 88 - 108 MHz, 200 kHz FM carrier spacing
- Peak-frequency deviation 75 kHz
- The FM radio broadcast receiver is superheterodyne type
  - IF of 10.7 MHz used

**FM Stereo**

- The left $L(t)$ and right $R(t)$ audio signals are added and subtracted to generate the sum $L + R$ and the difference signal $L - R$. The sum and difference signals are each preemphasized
- The $L - R$ signal DSB-SC modulates a 38-kHz subcarrier. 19 kHz pilot tone transmitted along with $L + R$ and $L - R$ signals
- $L + R$ signal used by mono FM receivers

FM Stereo (contd)

**FM stereo baseband frequency spectrum**
Black-and-White Image
- Formed by a large number of picture elements, called **pixels**
- **Raster scanning** encodes the light intensity of each pixel \( I(x,y) \) into electrical signal \( s(t) \)
- Scan each horizontal line and fast retrace to the next
- One complete set of lines makes a **frame** of the image.

Synchronization Signals
- **Synchronization (sync)** signals are added by the TV camera to allow reconstruction of the image at the receiver/monitor
  - Horizontal synchronizing pulse – at the end of each line
  - Horizontal blanking pulse – blank display during retrace
  - Vertical blanking – retrace to the start of a new frame

Video Signal
- Video picture sent as sequence of picture frames
  - Based on the principle that a rapid sequence of pictures can give the appearance of motion
  - Each picture digitized & compressed
  - Frame repetition rate
    - 10-30-60 frames/second depending on quality
    - NTSC system uses 525 lines/frame and displays 30 frames/second
  - **Uses interlacing.** A frame is divided into two **fields** that consist of the odd and even lines, respectively
    - Odd and even fields are displayed in alternation