Lecture 08: DSB-SC Detection

COSTAS loop demodulation of DSB-SC

Reference – 01: Rodger E. Ziemer, William H. Tranter-Principles of Communications, 7th Edition_ Systems, Modulation, and Noise-Wiley (2014)

4.3.4 Costas PLLs

We have seen that systems utilizing feedback can be used to demodulate angle-modulated carriers. A feedback system also can be used to generate the coherent demodulation carrier necessary for the demodulation of DSB signals. One system that accomplishes this is the Costas PLL illustrated in Figure 4.26. The input to the loop is the assumed DSB signal

$$x_r(t) = m(t)\cos(2\pi f_c t) \tag{4.145}$$

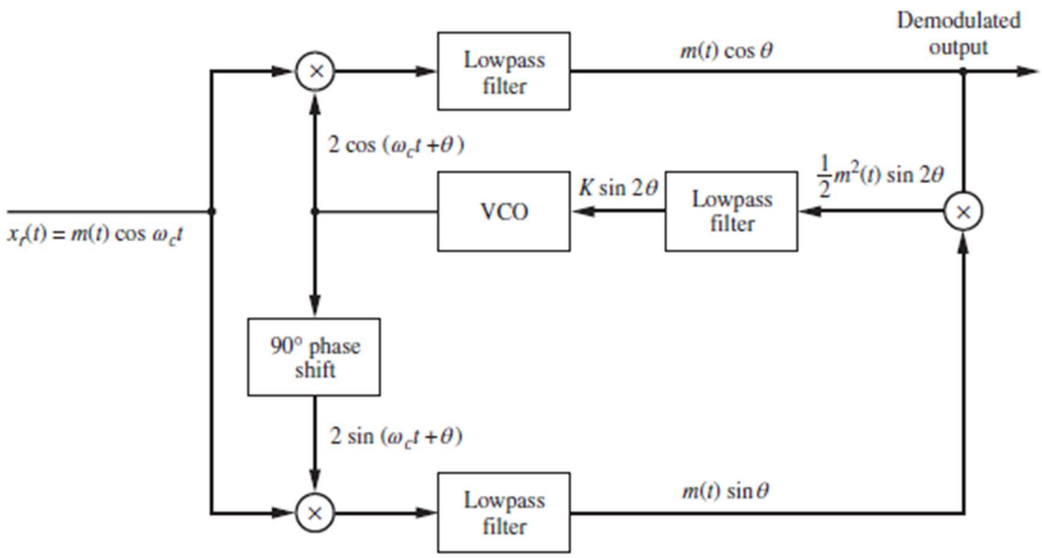


Figure 4.26
Costas phase-locked loop.

The signals at the various points within the loop are easily derived from the assumed input and VCO output and are included in Figure 4.26. The lowpass filter preceding the VCO is assumed to have sufficiently small bandwidth so that the output is approximately $K\sin(2\theta)$, essentially the DC value of the filter input. This signal drives the VCO such that θ is reduced. For sufficiently small θ, the output of the top lowpass filter is the demodulated output, and the output of the lower filter is negligible. We will later see in that the Costas PLL is useful in the implementation of digital receivers.

Reference – 02: Communication System Bruce Carlson 4ed

Synchronous Detection and Frequency Synthesizers

The lock-in ability of a PLL makes it ideally suited to systems that have a pilot carrier for synchronous detection. Rather than attempting to filter the pilot out of the accompanying modulated waveform, the augmented PLL circuit in Fig. 7.3–3 can be used to generate a sinusoid synchronized with the pilot. To minimize clutter here, we've lumped the phase comparator, lowpass filter, and amplifier into a phase discriminator (PD) and we've assumed unity sinusoidal amplitudes throughout.

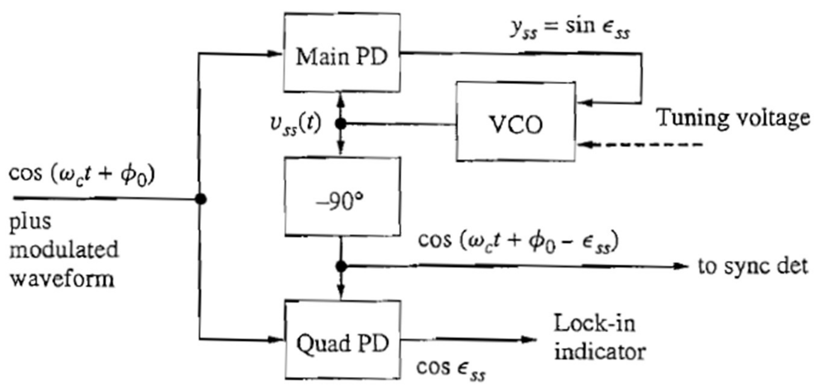


Figure 7.3–3 PLL pilot filter with two phase discriminators (PD).

Initial adjustment of the tuning voltage brings the VCO frequency close to f_c and $\epsilon_{ss} \approx 0$, a condition sensed by the quadrature phase discriminator and displayed by the lock-in indicator. Thereafter, the PLL automatically tracks any phase or frequency drift in the pilot, and the phase-shifted VCO output provides the LO signal needed for the synchronous detector. Thus, the whole unit acts as a *narrowband pilot filter* with a virtually noiseless output.

Incidentally, a setup like Fig. 7.3–3 can be used to *search* for a signal at some unknown frequency. You disconnect the VCO control voltage and apply a ramp generator to sweep the VCO frequency until the lock-in indicator shows that a signal has been found. Some radio scanners employ an automated version of this procedure.

For *synchronous detection* of DSB without a transmitted pilot, Costas invented the PLL system in Fig. 7.3–4. The modulated DSB waveform $x(t)\cos\omega_c t$ with bandwidth $2W$ is applied to a pair of phase discriminators whose outputs are proportional to $x(t)\sin\epsilon_{ss}$ and $x(t)\cos\epsilon_{ss}$. Multiplication and integration over $T \gg 1/W$ produces the VCO control voltage

$$y_{ss} \approx T\langle x^2(t)\rangle \sin\epsilon_{ss}\cos\epsilon_{ss} = \frac{T}{2} S_x \sin 2\epsilon_{ss}$$

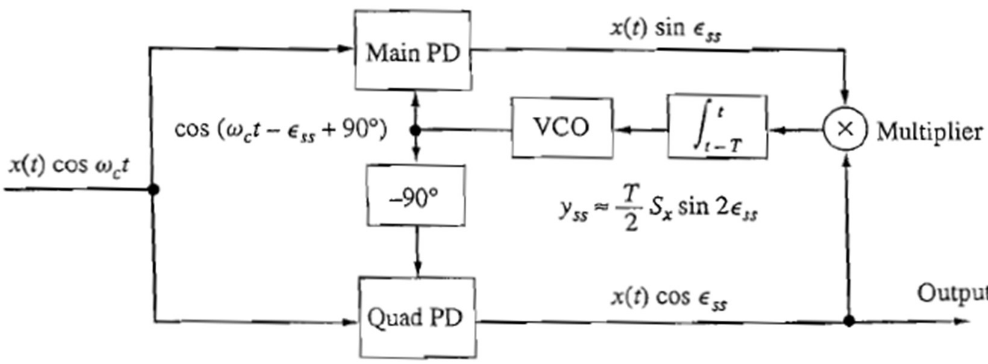


Figure 7.3–4 Costas PLL system for synchronous detection.

If $\Delta f \approx 0$, the PLL locks with $\epsilon_{ss} \approx 0$ and the output of the quadrature discriminator is proportional to the demodulated message $x(t)$. Of course the loop loses lock if $x(t) = 0$ for an extended interval.

The *frequency-offset loop* in Fig. 7.3–5 translates the input frequency (and phase) by an amount equal to that of an auxiliary oscillator. The intended output frequency is now $f_c + f_1$, so the free-running frequency of the VCO must be

$$f_v = (f_c + f_1) - \Delta f \approx f_c + f_1$$

The oscillator and VCO outputs are mixed and filtered to obtain the difference-frequency signal $\cos[\theta_v(t) - (\omega_1 t + \phi_1)]$ applied to the phase discriminator. Under locked conditions with $\epsilon_{ss} \approx 0$, the instantaneous angles at the input to the discriminator will differ by 90°. Hence, $\theta_v(t) - (\omega_1 t + \phi_1) = \omega_c t + \phi_0 + 90°$, and the VCO produces $\cos[(\omega_c + \omega_1)t + \phi_0 + \phi_1 + 90°]$.

By likewise equating instantaneous angles, you can confirm that Fig. 7.3–6 performs *frequency multiplication.* Like the frequency multiplier discussed in Sect. 5.2, this unit multiplies the instantaneous angle of the input by a factor of n. However, it does so with the help of a *frequency divider* which is easily implemented using a *digital counter.* Commercially available divide-by-n counters allow you to select any integer value for n from 1 to 10 or even higher. When such a counter is inserted in a PLL, you have an *adjustable* frequency multiplier.

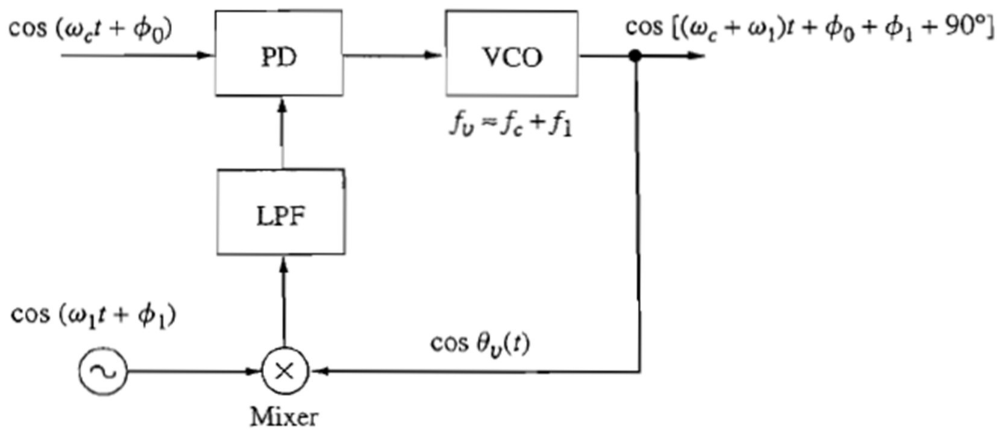


Figure 7.3–5 Frequency-offset loop.

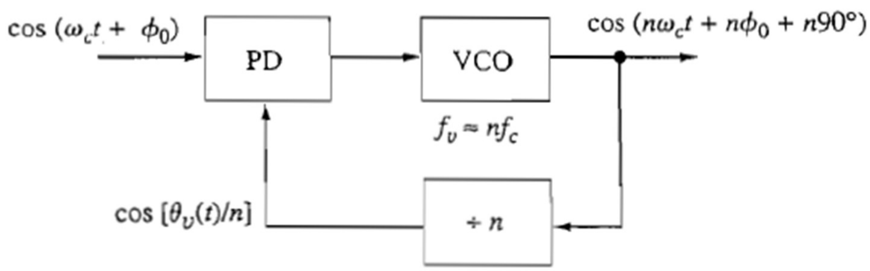


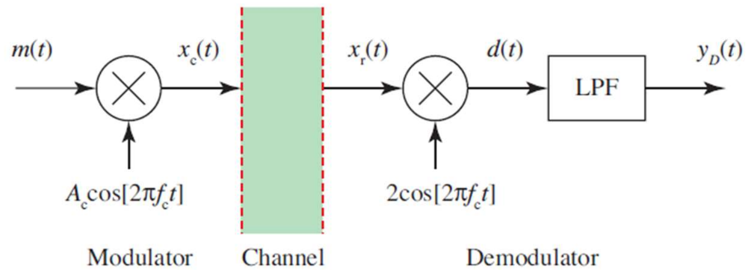
Figure 7.3–6 PLL frequency multiplier.

A *frequency synthesizer* starts with the output of one crystal-controlled master oscillator; various other frequencies are synthesized therefrom by combinations of frequency division, multiplication, and translation. Thus, all resulting frequencies are stabilized by and synchronized with the master oscillator. General-purpose laboratory synthesizers incorporate additional refinements and have rather complicated diagrams. So we'll illustrate the principles of frequency synthesis by an example.

Reference – 03: ECE 5625/4625 Communication Systems Notes by Dr Woickert

Coherent Demodulation

- The received signal is multiplied by the signal $2\cos(2\pi f_c t)$, which is synchronous with the transmitter carrier

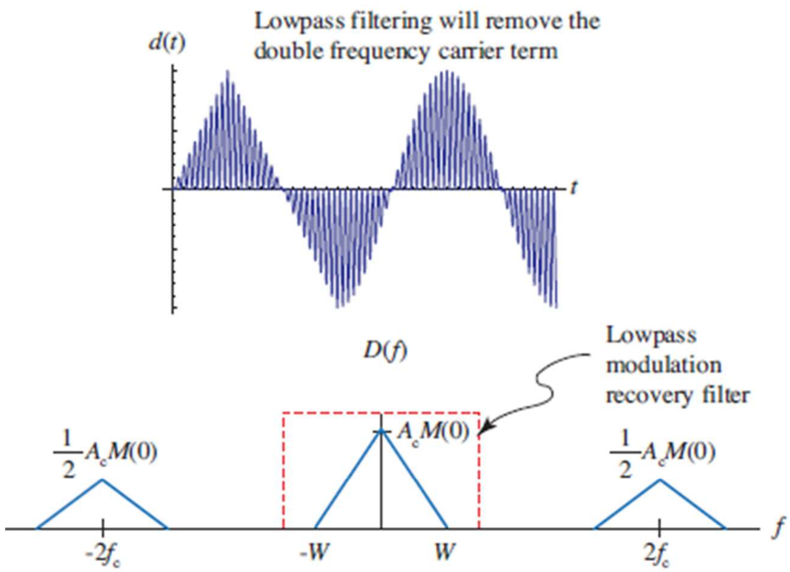


- For an ideal channel $x_r(t) = x_c(t)$, so

$$\begin{aligned} d(t) &= \left[A_c m(t)\cos(2\pi f_c t)\right] 2\cos(2\pi f_c t) \\ &= A_c m(t) + A_c m(t)\cos(2\pi(2f_c)t) \end{aligned}$$

  where we have used the trig identity $2\cos^2 x = 1 + \cos 2x$

- The waveform and spectra of $d(t)$ is shown below (assuming $m(t)$ has a triangular spectrum in $D(f)$)



Waveform and spectrum of $d(t)$

- Typically the carrier frequency is much greater than the message bandwidth W, so $m(t)$ can be recovered via lowpass filtering

- The scale factor A_c can be dealt with in downstream signal processing, e.g., an automatic gain control (AGC) amplifier

- Assuming an ideal lowpass filter, the only requirement is that the cutoff frequency be greater than W and less than $2f_c - W$
- The difficulty with this demodulator is the need for a coherent carrier reference
- To see how critical this is to demodulation of $m(t)$ suppose that the reference signal is of the form

$$c(t) = 2\cos[2\pi f_c t + \theta(t)]$$

  where $\theta(t)$ is a time-varying phase error

- With the imperfect carrier reference signal

$$d(t) = A_c m(t)\cos\theta(t) + A_c m(t)\cos[2\pi f_c t + \theta(t)]$$
$$y_D(t) = m(t)\cos\theta(t)$$

- Suppose that $\theta(t)$ is a constant or slowly varying, then the $\cos\theta(t)$ appears as a fixed or time varying attenuation factor
- Even a slowly varying attenuation can be very detrimental from a distortion standpoint
  - If say $\theta(t) = \Delta f t$ and $m(t) = \cos(2\pi f_m t)$, then

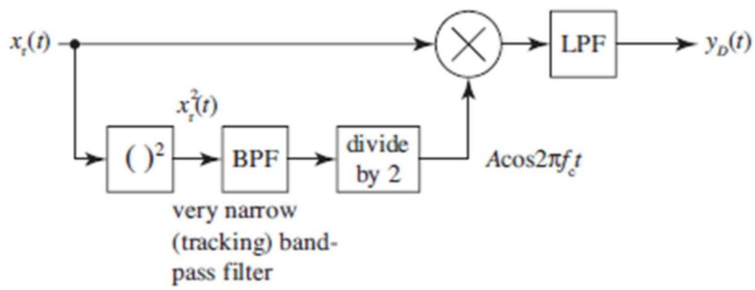
$$y_D(t) = \frac{1}{2}[\cos[2\pi(f_m - \Delta f)t] + \cos[2\pi(f_m + \Delta f)t]]$$

    which is the sum of two tones

- Being able to generate a coherent local reference is also a practical manner

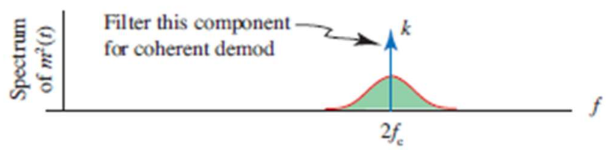
- One scheme is to simply square the received DSB signal

$$
\begin{aligned}
x_r^2(t) &= A_c^2 m^2(t)\cos^2(2\pi f_c t)\\
&= \frac{1}{2}A_c^2 m^2(t) + \frac{1}{2}A_c^2 m^2(t)\cos[2\pi(2f_c)t]
\end{aligned}
$$



Carrier recovery concept using signal squaring

- Assuming that $m^2(t)$ has a nonzero DC value, then the double frequency term will have a spectral line at $2f_c$ which can be divided by two following filtering by a narrowband bandpass filter, i.e., $\mathcal{F}\{m^2(t)\} = k\delta(f) + \cdots$



- Note that unless $m(t)$ has a DC component, $X_c(f)$ will not contain a carrier term (read $\delta(f \pm f_c)$), thus DSB is also called a *suppressed carrier* scheme