CARRIER-DETECTOR EXTENDS ACQUISITION RANGE OF BPSK

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Abstract

The capture and tracking range of a Costas type BPSK (QPSK) demodulator is limited by its loop bandwidth, which is determined by the synchronization threshold at different Signal-to-Noise Ratio. A frequency locked scheme, which consists of a carrier detector and digital control network, has been developed to work in parallel with the Costas loop, so the BPSK (QPSK) demodulator would be automatically led to its pull-in range at low Signal-to-Noise Ratio. The probabilities of false dismissal for signal and false alarm signal absent are analyzed.

Performance Limitations of a Phase Locked Loop (PLL) Circuit

There are several kinds of carrier-recovery circuits used in BPSK (QPSK) demodulators. Let's consider here only Costas-Loop BPSK demodulator, which has the following equivalent phase-locked loop:

Closed Loop transfer function, $G(s) = \dfrac{K_\phi K_v (s+a)}{s^2 + K_\phi K_v s + K_\phi K_v a}$

it's loop filter transfer function, $H(s) = \dfrac{s+a}{s}$

$K_\phi = \dfrac{A\ Km}{4}$, is the Phase-Comparator constant, in $^{Volt}/_{Rad}$

A is rms of voltage of the input signal fed to in-phase and quadrature product demodulator. K_m is the multiplier constant in volt/volt. K_v is the VCO sensitivity in rad/volt.

Then, the loop nature frequency, $\omega_a = \sqrt{K_\phi K_v a}$ rad/sec., and the damping

factor $\zeta = \sqrt{\dfrac{K_\phi K_v}{4a}}$

The minimum noise bandwidth occurs at $\zeta = 0.5$, $B_N = \frac{\omega_n}{2}$, usually, ζ is set to 0.707, which results in a noise bandwidth B_N, only 6% (0.25dB noise power) greater than the minimum noise bandwidth.

When the initial frequency off-set is within the capture range, the synchronization-acquisition time is approximately the sum of frequency-lock time, T_{FL}, and phase-lock time, T_{PL}.

$$T_{FL} \simeq 4 \left(\frac{\Delta f^2}{B_N^3} \right); \quad T_{PL} \simeq 1.3 / B_N$$

B_N is the equivalent noise bandwidth of PLL.

Viterbi[1] has shown that the region of "frequency lock", $\Delta\omega$, within which the loop stops skipping cycles after pull into a diminishing phase error is approximately given by

$$\Delta\omega \simeq 2\omega_n (\zeta + 0.6), \ \zeta > 0.3$$

With noise present, for a fixed loop width, the synchronization loses lock at certain Signal-to-Noise Ratios, and this S/N is called synchronization threshold. The sync threshold is usually defined as the S/N at which probability of a clock slippage is 10^{-6}. From theoretical analysis, the sync threshold found is overly optimistic. A typical synchronization threshold chart from practical result is given in Figure 1.

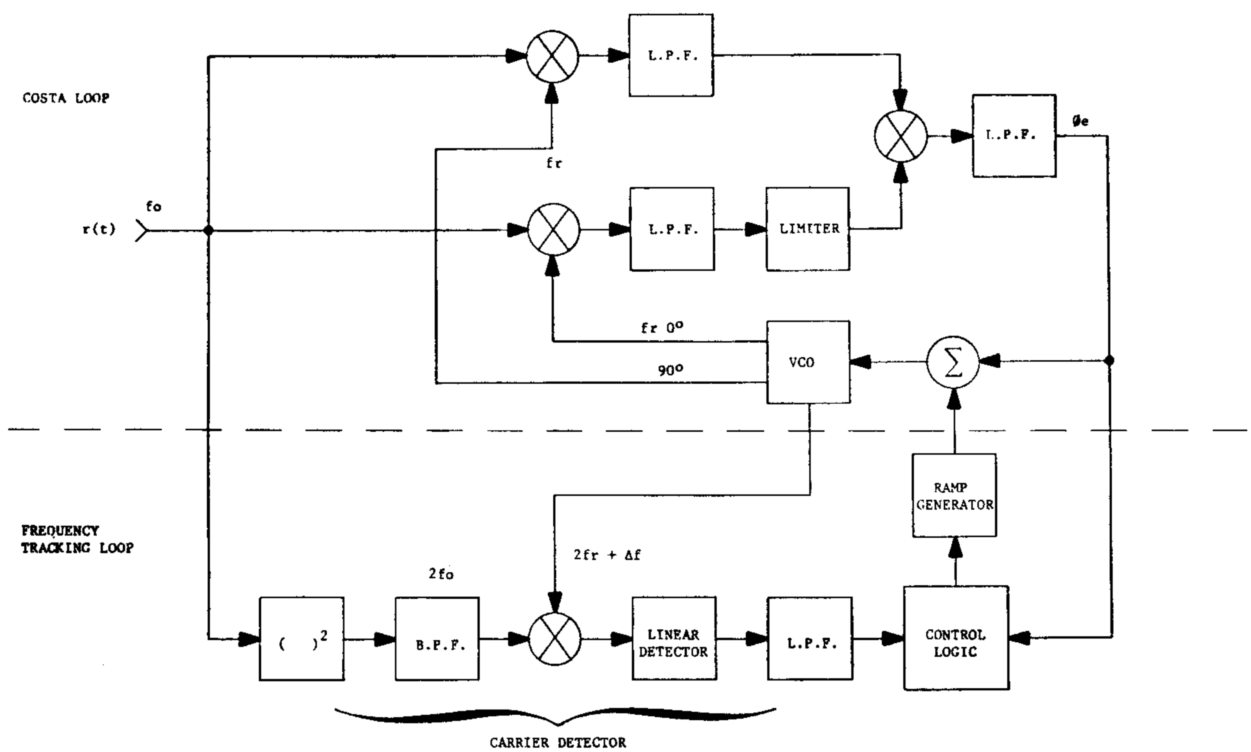
The loop bandwidth tends to be very narrow when the E_b/N_o specified goes to OdB. The capture range, $\Delta\omega$, in terms of the loop filter $H(j\omega)$, decreases as the low-pass filter time constant goes large, and the mean square phase jitter at the synchronizer's output is reduced.

In applications where wide capture range is not required, one can use a small time constant to obtain wider capture range, after the loop being in steady state, then switch the time constant to a large value, to have a narrower loop bandwidth for better synchronization. However, in applications where the carrier frequency uncertainty is very wide and searching time is important, switching loop filter time constant is not a practical solution.

FREQUENCY SEARCH AND ACQUISITION

A narrow loop bandwidth is essential for a high performance BPSK demodulator. Here, we present a scheme for the above described PLL. It locates the offset carrier, tracks with it, while the loop bandwidth is still retained very narrow.

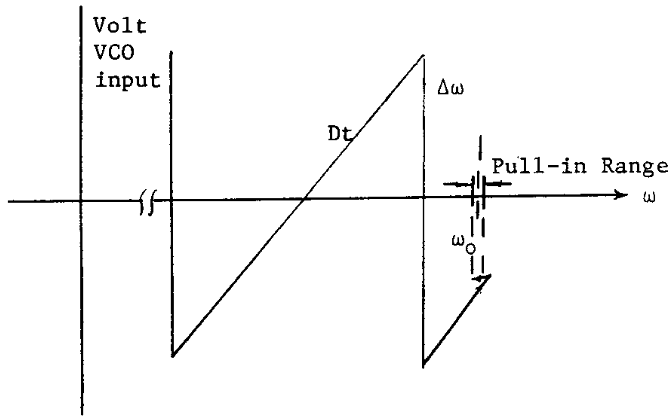
Acquisition behavior is described in terms of carrier-tracking application where the center frequency of the carrier is fixed or changing at a rate that is very slow compared to the closed-loop dynamics of the phase-locked loop, such as the doppler effect due to satellite's movement.

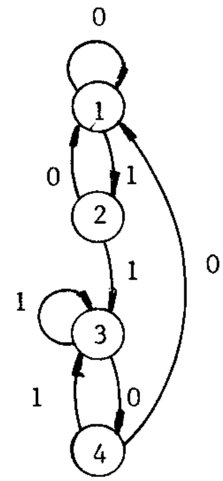


The simplified block diagram shows the inter-connections between the two loops. The frequency tracking loop has three parts -- a non-coherent detector, a ramp-generator and the decision logic. The ramp drives VCO sweeping the entire frequency range of carrier uncertainty; when carrier is detected, the control logic stops the sweeping and changes integrator's time constant to very large, then directs the ramp going up or down according to the phase error signal's polarity.

The Ramp Generator consists of two independent ramp generators: one is a stair-case generator, incremented by a digital counter, its output can be held at the same level for any long period of time; the other is a very slow integrator.

The control logic is a simple four-state state machine. The whole operation is self-explained as in the flow chart, Figure 2.





When the stair-case generator stops,
VCO output frequency is a little bit overshot.
Therefore, the slow-ramp generator sweeps back
until the Pull-in range of the Costa Loop is met.

State Transistion Diagram
See Fig 2, Flow-Chart for
State Definition

Non-coherent Detector

The non-coherent detector consists of:

a squaring circuit, which removes the modulation ±A, and creates a line component in the spectrum at double the carrier frequency $2\omega_o$;
a narrow band-pass filter, which removes all the signals except the $2\omega_o$;
a down-converter, which shifts the signal to $2\omega_o$, the off-set frequency from Costas loop's local oscillator;
a linear detector and single pole RC low-pass filter, which outputs the detected d.c. signal, Z, with noise to a comparator. When the d.c. level is higher than the pre-set threshold, Z_T, means that signal is present. The decision logic branches to next state, as the State Diagram shows.

When the Costas loop is frequency-locked, carrier detector output has the peak power of $2\omega_o$. The DSI 7132 BPSK demodulator accepts an input carrier frequency of 70MHz. The frequency tracking loop extends the carrier uncertainty to ±250KHz; we have achieved a maximum acquisition time less than 5 seconds at OdB E_b /N_o. The fast search rate is 100KHz/second; mean phase-lock time is less than 0.2 seconds, while the loop bandwidth is retained at about 2KHz.

Probability of False Alarm

Let the input phase-modulated signal added with white noise be first filtered by a rectangular band pass filter (SAW) of bandwidth W. The signal with noise is given as:

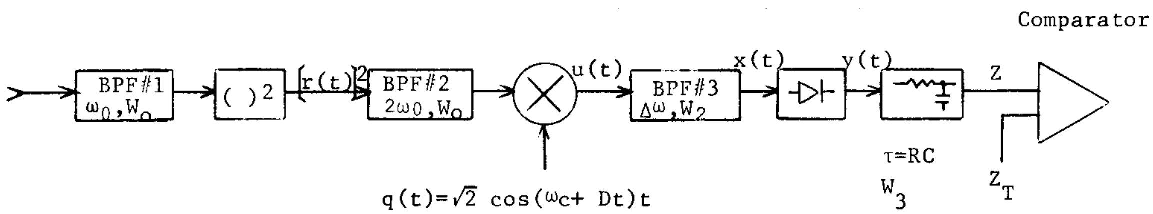
$r(t) = \sqrt{2}\, A\, s(t) \sin(\omega_o t+\theta) + n(t)$

$s(t) = \pm 1$

n(t) has band limited white spectral density (one side) N_o, the signal has total power of $P_s = A^2$

Then the squaring circuit output is:

$[r(t)]2 = A^2 - A^2 \mathrm{Cos}2(\omega_o t+\theta) + 2\sqrt{2}\, A \,\mathrm{Sin}(\omega_o t + \theta)\, n(t) + n^2(t)$



The bandpass filter following the square law device filters out all d.c.

The power in the signal term at frequency $2\omega_o$ is $\dfrac{A^4}{2} = \dfrac{Ps^2}{2}$

the noise power spectral density at $2\omega_o$ is approximately equal to:

$$N_o' \simeq 2A^2 N_o + N_o^2 W_o$$

which has an amplitude spectrum decreasing linearly from $2\omega_o$ to $2\omega_o \pm W_o$; because of $W_2 << W_o$, we assume that the noise in the spectrum region of $2\omega_o$ is still white noise. The Signal-to-Noise ratio at the output of band pass filter No. 2 is:

$$(SNR)_{2\omega_o} = \frac{A^4/2}{N_o' W_1} = \frac{P_s^2/2}{W_1[2A^2 N_o + N_o^2 W_o]} = \frac{P_s}{4 N_o W_1} \quad \frac{1}{1 + N_o/_{2E_b}}$$

let $\omega_o' = 2\omega_o$, $P_s' = \dfrac{P_s^2}{2}$

$q(t) = \sqrt{2}\cos(\omega_c + Dt)t$, the sweeping oscillator signal. The linear sweeping rate: $D = \Delta\omega/_T$, that is, the signal stays within the pull-in range $\Delta\omega$ for T seconds.

At the mixer output,

$$u(t) = \sqrt{P_s'}\ \mathrm{Cos}\ (\Delta\omega_o + Dt)t + \frac{N'(t)}{\sqrt{2}} \cos\left[(\Delta\omega_o + Dt)t + \phi_n(t)\right]$$

W_2 is narrow, with respect to the signal bandwidth, but is wide compared to the observation time. As the VCO sweeps, the signal falls in the $\Delta\omega_o$ pass band, (the Costas loop in the pull-in range $\Delta\omega$), the linear detector has an input

$$x(t) = \sqrt{P_s'} \cos\ (\Delta\omega_o t) + N'(t) \cos\left[\Delta\omega_o t + \phi_n'\ (t)\right].$$

That is, when the signal is present in the $\Delta\omega_o$ Awo, the detector output

$$y(t) = x^2(t)_{\ \mathrm{LowPass}} \simeq \frac{P_s'}{2} + \frac{N'^2(t)}{2} + \sqrt{P_s' N'}(t)$$

when signal is outside of $\Delta\omega$ pass band, or is absent, then $y(t) = \frac{N'^2(t)}{2}$

This y(t) has a Rayleigh probability density.

The low pass detection filter is a single pole RC with a time constant T and noise bandwidth W_3, $T = \frac{W_3}{2}$, and $W_3 \ll W_2$ so that the low pass filter output is approximately Gaussian. When the signal is inside the $\Delta\omega$ for T seconds, the mean output of the low pass filter is:

$$\overline{E\ [z(T)]} = \frac{P_s'}{2}(1-\alpha) + \frac{N_o' W_2}{2} \quad \text{signal present}$$

$$= \frac{N_o' W_2}{2} \quad \text{signal absent}$$

$$\alpha = e^{-\frac{T}{\tau}}$$

the variance of the low-pass filter output is:

$$\sigma_z^2 = P_s'\ (1-\alpha^2)\ \frac{N_o W_3}{2} + \left(\frac{N_o}{2}\right)^2 \frac{W_2 W_3}{2} \quad \text{signal present}$$

$$= \left(\frac{N_o}{2}\right)^2 \frac{W_2 W_3}{2} \quad \text{signal absent}$$

Thus, for a threshold Z_T, the probability of false dismissal for signal present is:

$$\mathcal{P}_d(Z<Z_t)\ \text{Sig Present} = \frac{1}{2}\,\text{erfc}\ \frac{1-\alpha+\frac{1}{(SNR)_{W_2}}-Z_T}{\sqrt{\left[2(1-\alpha^2)+\frac{1}{(SNR)_{W_2}}\right]\frac{1}{(SNR)_{W_2}}\ \frac{W_3}{W_2}}}$$

$$\text{where } (SNR)_{W_2} = \frac{P_s'}{W_2 N_o'} = \frac{P_s^2}{2W_2(2A^2N_o+N_o^2W_o)}$$

The probability of false alarm is:

$$\mathcal{P}_a(Z>Z_T)_{\text{Sig Absent}} = \frac{1}{2}\,\text{erfc}\ \frac{Z_T-\frac{1}{(SNR)_{W_2}}}{\sqrt{\frac{4W_3}{W_2}\ \frac{1}{(SNR)_{W_2}}}}$$

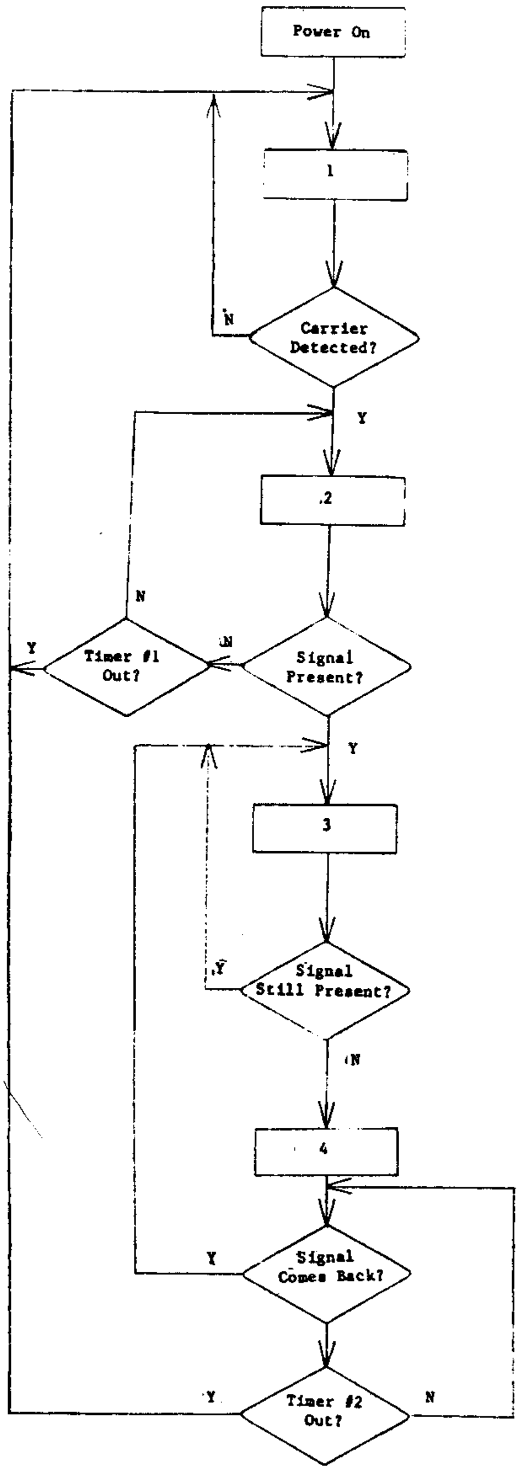
Following a narrow band pass filter, the squaring circuit is to remove the sideband of the phase-modulated signal. Also, it causes envelope modulation at each phase reversal. Thus, there would be a pair of line frequency components at $2\omega_0 \pm \frac{2\pi}{T_d}$, where T_d is the bit period. T_d is the bit period. These bit frequency line components can cause false lock if the VCO sweeps wider than $\pm \frac{1}{T_d}$ Hz. This is the only limitation for this scheme on the searching range.

References:

[1] "Acquisition and Tracking Behavior of Phase-locked Loops." by Viterbi, A.J., Proc. Symposium on Active Network, New York, Polytechnic Press, 1960 = 583-619,b

[2] "Digital Communications by Satellite", by Spilker, J.J.Jr., Prentice-Hall, 1977

OPERATION FLOW CHART



1. Fast Search:	VCO sweeping at 100KHz/sec rate.
2. Slower Search:	Fast-Ramp-Generator is Hold; Slow-Ramp-Generator starts. VCO sweeps very slowly; while "Timer #1" is enabled.
3. Track:	Phase comparator and Slow-Ramp (up-down) work together as closed loop PLL.
4. Wait:	Restart slow search. "Timer #2" is enabled; and the indication light on front panel starts blinking.

FIG 2

Hardware Implement of Frequency Tracking Loop on DSI 7132 BPSK Demodulator

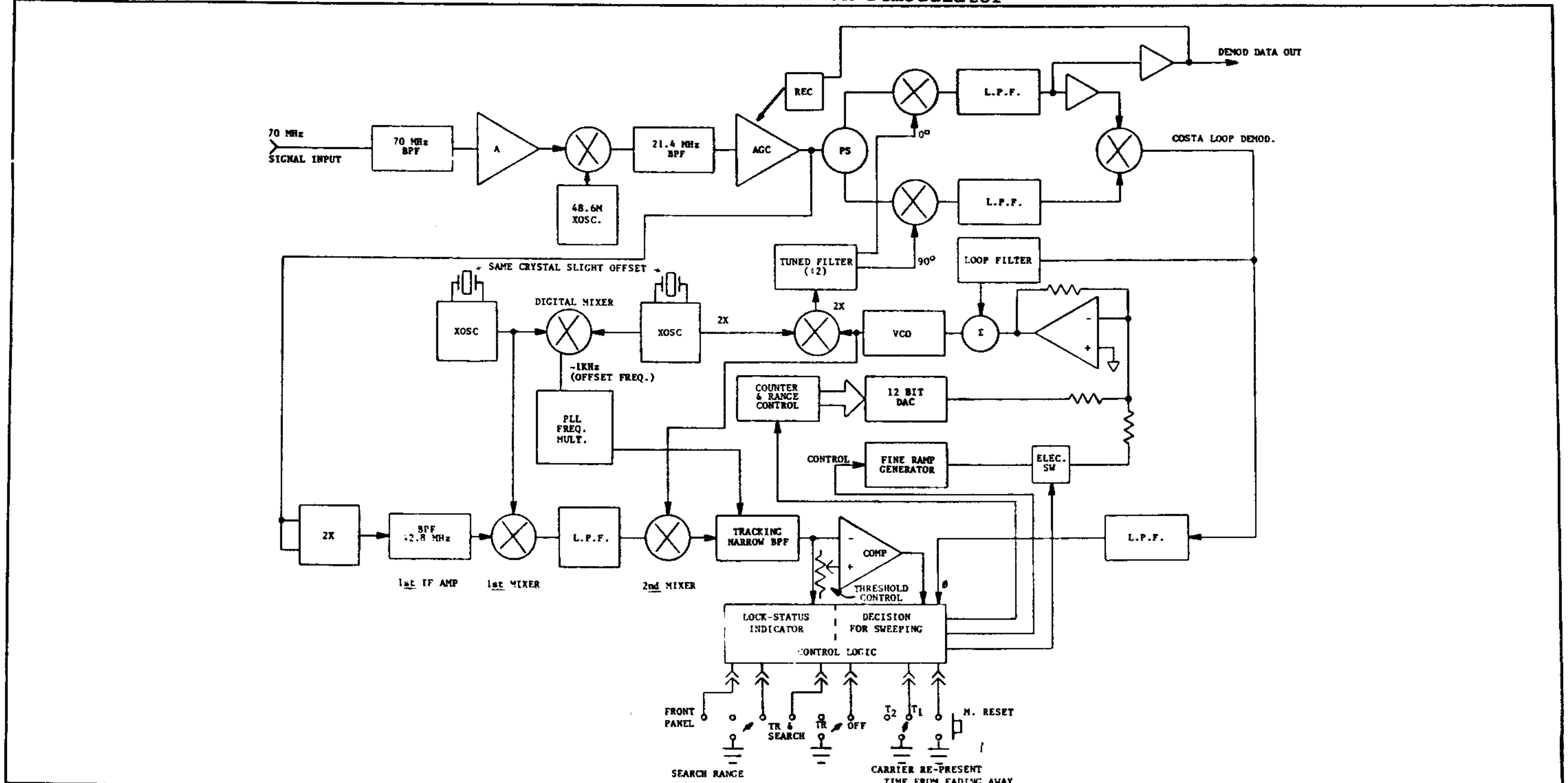


FIG 3