

PHASE LOCKED LOOP USING DELAY COMPENSATION TECHNIQUES

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FIELD OF THE INVENTION

The present invention relates to phase locked loop topology including and more particularly the phase locked loop technology for compensating for the effects of loop latency in a read channel.

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BACKGROUND OF THE INVENTION

The design of phase locked loops for hard disk drives (HDD) data recovery has been used by implementing decision directed algorithms to adjust the frequency and phase of the recovered clock in order to optimize the sampling instants (timing instants) of the received signals. These algorithms which, by their very nature, depend on a
15 reliable estimate (decision) of the current sample, have grown in complexity as the read channel signal processing has moved to higher partial response polynomials. One way of obtaining reliable phase error estimates relies on obtaining decision (binning) information from the Viterbi detector. Because of the complexity associated with this block, these topologies are characterized by latencies, usually symbolized by the term
20 mD, which can be easily in excess of 20 clock cycles. Figure 1 illustrates one such circuit.

The excessive latency greatly reduces the region in which the PLL is stable. As an example, Figures 2 and 3 show how one such region varies as the latency increases
25 from one clock cycle (ideal) to 19 clock cycles (representative of a medium latency digital PLL). In practical terms, besides an increase in the complexity of the system model, the larger the latency, relative to the time scale of the dynamics of the process,



the loop gain reduction needs to be larger. Under most circumstances, this causes poor performance and sluggish response within the PLL.

Loop latency also has a significant impact on the design of the time base
s generator, which is a circuit that generates the reference clock for the PLL.

Thus, there has been a need for topology which decouples the clock recovery circuit performance from the loop latency while maintaining reliability in the phase detection process.

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SUMMARY OF THE INVENTION

The present invention provides a circuit that eliminates the latency associated with external delays. A feedback loop is provided to eliminate these delays. The
5 present invention provides a zero steady state error in the presence of offsets between the frequency of the input signal and the frequency of the reference clock.

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BRIEF DESCRIPTION OF THE DRAWINGS

Figure 1 illustrates a PLL circuit;

5 Figure 2 illustrates a stability region as a function of latency;

Figure 3 illustrates another stability region as a function of latency;

10 Figure 4 illustrates a circuit diagram of the present invention;

Figure 5 illustrates another circuit diagram of the present invention;

Figure 6 illustrates another circuit diagram of the present invention;

15 Figure 7 illustrates another circuit diagram of the present invention;

Figure 8 illustrates another circuit diagram of the present invention;

Figure 9 illustrates another circuit diagram of the present invention;

20 Figure 10 illustrates another circuit diagram of the present invention;

Figure 11 illustrates another circuit diagram of the present invention;

25 Figure 12 illustrates a stability region as a function of latency;

Figure 13 shows another stability region as a function of latency;

30 Figure 14 shows another stability region as a function of latency;



Figure 15 shows comparisons of the system dynamics in response to a phase step;

Figure 16 shows comparisons of the system dynamics in response to a
5 frequency step;

Figure 17 shows comparisons of the system dynamics in response to a frequency ramp;

10 Figure 18 shows graphs of the present invention;

Figure 19 is a side view of a disk drive system; and

Figure 20 is a top view of a disk drive system.

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DETAILED DESCRIPTION OF THE PRESENT INVENTION

The following invention is described with reference to figures in which similar or the same numbers represent the same or similar elements. While the invention is described in terms for achieving the invention's objectives, it can be appreciated by those skilled in the art that variations may be accomplished in view of these teachings without deviation from the spirit or scope of the invention.

Figures 19 and 20 show a side and top view, respectively, of the disk drive system designated by the general reference 100 within an enclosure 110. The disk drive system 100 includes a plurality of stacked magnetic recording disks 112 mounted to a spindle 114. The disks 112 may be conventional particulate or thin film recording disk or, in other embodiments, they may be liquid-bearing disks. The spindle 114 is attached to a spindle motor 116 which rotates the spindle 114 and disks 112. A chassis 120 is connected to the enclosure 110, providing stable mechanical support for the disk drive system. The spindle motor 116 and the actuator shaft 130 are attached to the chassis 120. A hub assembly 132 rotates about the actuator shaft 130 and supports a plurality of actuator arms 134. The stack of actuator arms 134 is sometimes referred to as a "comb." A rotary voice coil motor 140 is attached to chassis 120 and to a rear portion of the actuator arms 134.

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A plurality of head suspension assemblies 150 are attached to the actuator arms 134. A plurality of inductive transducer heads 152 are attached respectively to the suspension assemblies 150, each head 152 including at least one inductive write element. In addition thereto, each head 152 may also include an inductive read element or a MR (magneto-resistive) read element. The heads 152 are positioned proximate to the disks 112 by the suspension assemblies 150 so that during operation, the heads are in electromagnetic communication with the disks 112. The rotary voice coil motor 140 rotates the actuator arms 134 about the actuator shaft 130 in order to move the head suspension assemblies 150 to the desired radial position on disks 112.



A controller unit 160 provides overall control to the disk drive system 100, including rotation control of the disks 112 and position control of the heads 152. The controller unit 160 typically includes (not shown) a central processing unit (CPU), a memory unit and other digital circuitry, although it should be apparent that these aspects could also be enabled as hardware logic by one skilled in the computer arts. Controller unit 160 is connected to the actuator control/drive unit 166 which is in turn connected to the rotary voice coil motor 140. A host system 180, typically a computer system or personal computer (PC), is connected to the controller unit 160. The host system 180 may send digital data to the controller unit 160 to be stored on the disks, or it may request that digital data at a specified location be read from the disks 112 and sent back to the host system 180. A read/write channel 190 is coupled to receive and condition read and write signals generated by the controller unit 160 and communicate them to an arm electronics (AE) unit shown generally at 192 through a cut-away portion of the voice coil motor 140. The AE unit 192 includes a printed circuit board 193, or a flexible carrier, mounted on the actuator arms 134 or in close proximity thereto, and an AE module 194 mounted on the printed circuit board 193 or carrier that comprises circuitry preferably implemented in an integrated circuit (IC) chip including read drivers, write drivers, and associated control circuitry. The AE module 194 is coupled via connections in the printed circuit board to the read/write channel 190 and also to each read head and each write head in the plurality of heads 152. The read/write channel 190 includes the PLL circuit of the present invention.

The present invention cancels the effects that latency has on the timing loop of a PLL circuit and is based upon the concept that every time the PLL Vco phase is updated, it obeys the following equation (Figure 1):

$$\phi_{vco_n} = \phi_{vco_{n-1}} + \text{phUpdate}_{n-1}$$

(1)



On the other side, by definition of the phase error, Equation (2) follows.

$$\phi_{err_n} = \theta_{in_{n-1}} - \theta_{vco_{n-1}} = \theta_{in_{n-1}} - \phi_{vco_{n-m-1}} \quad (2)$$

5 One can compute exactly the amount by which the phase error would have to be reduced as a consequence of the update in the Vco phase. One way to eliminate the effects of latency is to remove the correction for latency from the next phase error term. This can be done by feeding back the Vco phase information and subtracting it from the phase error. We will call the results of this operation the delay compensated phase
10 error which is defined by Equation (3).

$$\lambda_n = \phi_{err_{n-1}} - \phi_{vco_{n-1}} = \theta_{in_{n-2}} - \phi_{vco_{n-m-2}} - \phi_{vco_{n-1}} \quad (3)$$

and has three distinct components as illustrated in Equation (3a).

$$\begin{aligned} \theta_{in_{n-1}} & , \quad \text{input related phase changes} \\ \phi_{vco_{n-m-1}} & , \quad \text{delayed vco phase term} \\ \phi_{vco_{n-1}} & \quad \text{delay compensation term} \end{aligned} \quad (3a)$$

20 Equation (3) implies that the PLL loop would process the same correction twice, once at time index $n-1$ and once at time index $n-m-2$. Since this is not desirable, it is more desirable to eliminate the $\phi_{vco_{n-m-2}}$ term. This operation can be accomplished by redefining λ as illustrated in Equation (4).

$$\lambda_n = \phi_{err_{n-1}} - \phi_{vco_{n-1}} = \theta_{in_{n-2}} - \phi_{vco_{n-m-2}} - \phi_{vco_{n-1}} = \theta_{in_{n-2}} - \phi_{vco_{n-1}} \quad (4)$$



The implementation of this concept is shown in Figure 4. θ_n is input to adder circuit 402. Additionally, θ_{vco} is input to the negative input of adder circuit 402. The output signal ϕ of adder circuit 402 is input to adder circuit 404. In addition, signal $\gamma\psi$ is input to adder circuit 404. Output from adder circuit 404 is the signal λ which is input both to multiplier 406 and multiplier 408. The multiplier 406 amplifies λ by α and is output to adder circuit 410. The multiplier circuit 408 multiplies λ by ρ and outputs $\rho\lambda$ to the adder circuit 412. The output of adder circuit 412 is delayed by delay circuit 414 and input to adder circuit 412. The output of adder circuit 412 is additionally input to adder circuit 410. The output of adder circuit 410 is signal $phUpdate_{n-1}$ and input to adder circuit 416. The output of adder circuit 416 is ϕ_{vco} which is input to delay circuit 418. The output of delay circuit 418 is input to adder circuit 416. The output of adder circuit 416 ϕ_{vco} is input both to adder circuit 420 and delay circuit 422. The delay circuit 422 represents the delay associated with the sampling (A/D), equalization (FIR) and phase detector circuits. The output of delay circuit 422 is input to adder circuit 420 and also as θ_{vco} to adder circuit 402. The output of adder circuit 420 $\gamma\psi$ is input to adder circuit 404.

The proportional branch 432 includes multiplier 406, and the integral branch 434 includes multiplier 408, adder 412 and delay 414. Adder 412 in conjunction with delay 414 is also an integral accumulator. The Vco accumulator 436 includes adder 416 and delay 418. The PII filter 430 includes the proportional branch 432 and the integral branch 434.

This topology accomplishes the stated advantage of making the performance of the overall timing loop strongly decoupled from m , depending solely on the delay associated with the loop filter, the Vco accumulator and the delay compensation loop. With reference to Figure 4, we can write the delay compensated phase error signal λ_n in Equation (5).



$$\lambda_n = \phi_{err_{n-1}} - \gamma \psi_{n-1} = (\theta_{in_{n-2}} - \theta_{vco_{n-2}}) - \gamma (\phi_{vco_{n-2}} - \theta_{vco_{n-2}}) = (\theta_{in_{n-2}} - \gamma \phi_{vco_{n-2}}) - (\theta_{vco_{n-2}} - \gamma \theta_{vco_{n-2}}) \quad (5)$$

Here, ϕ_{err} is the phase error, ψ_n is the delay compensation feedback. The delay compensation gain γ is a parameter which allows to turn on and turn off the effects of the delay compensating network. Since θ_{vco} is the delayed version of ϕ_{vco} , we can write Equation (6):

$$\theta_{vco_n} = \phi_{vco_{n-m}} \quad (6)$$

where m is the portion of the overall loop latency due to the sampling network (e.g. A/D), equalization circuits (e.g. FIR) and phase detector (inclusive of Viterbi processing latency) of the read channel 190. θ_{vco} is the phase of the sampling clock. The fundamental Vco update equation is shown in Equation (7):

$$\phi_{vco_n} = \phi_{vco_{n-1}} + \text{phUpdate}_{n-1} \quad (7)$$

Combining (6) and (7) leads to Equation (8):

$$\theta_{vco_n} = \theta_{vco_{n-1}} + \text{phUpdate}_{n-m} \quad (8)$$

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For a traditional PI filter, we have Equation (9):

$$\text{phUpdate}_n = \alpha \lambda_{n-1} + \rho \sum_{j=1}^{n-1} \lambda_j \quad (9)$$

25 Combining (8) and (9) yields Equation (10):



$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \alpha (\lambda_{n-m-3} - \lambda_{n-m-4}) + \rho \lambda_{n-m-4} \quad (10)$$

and substituting (5) into (10) yields Equation (11):

$$\begin{aligned} \theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = & \alpha ((\theta in_{n-m-5} - \gamma \theta vco_{n-m-5}) - (\theta vco_{n-m-5} - \gamma \theta vco_{n-m-6}) - \\ & (\theta in_{n-m-6} - \gamma \theta vco_{n-m-6}) + (\theta vco_{n-m-6} - \gamma \theta vco_{n-m-7})) + \\ & \rho ((\theta in_{n-m-6} - \gamma \theta vco_{n-m-6}) - (\theta vco_{n-m-6} - \gamma \theta vco_{n-m-7})) \end{aligned} \quad (11)$$

This provides a baseline for comparison with the present invention. We discuss the uncompensated case for which $\gamma = 0$.

Open Loop Gain

From Equations (5) and (10) we obtain Equation (12):

$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \alpha (\phi err_{n-m-4} - \phi err_{n-m-5}) + \rho \phi err_{n-m-5} \quad (12)$$

Taking the z-transform of (12), we obtain an open loop transfer function in Equation (13):

$$OLGain_0 = z^{-m} \frac{\alpha (z^{-5} - z^{-6}) + \rho z^{-6}}{1 - 2z^{-1} + z^{-2}} = \frac{z^{-4-m} ((-1+z)\alpha + \rho)}{(1+z)^2} \quad (13)$$

As expected, the open loop transfer functions contain a z^{-m} factor associated with loop latency.

Closed Loop Gain



The closed loop transfer function can be obtained directly from Equation (11) which, for $\gamma = 0$, simplifies to Equation (14):

$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \alpha ((\theta in_{n-m-5} - \theta vco_{n-m-5}) - (\theta in_{n-m-6} - \theta vco_{n-m-6})) + \rho (\theta in_{n-m-6} - \theta vco_{n-m-6}) \quad (14)$$

Equation (14) can be manipulated to yield Equation (15):

$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} + \alpha (\theta vco_{n-m-5} - \theta vco_{n-m-6}) + \rho \theta vco_{n-m-6} = \alpha (\theta in_{n-m-5} - \theta in_{n-m-6}) + \rho \theta in_{n-m-6} \quad (15)$$

Taking the z-transform of (15), the following closed loop transfer function in Equation (16) is obtained:

$$ClGain_0 = z^{-m} \frac{\alpha (z^{-5} - z^{-6}) + \rho z^{-6}}{1 - 2z^{-1} + z^{-2} + z^{-m} (\alpha (z^{-5} - z^{-6}) + \rho z^{-6})} \cdot \frac{(-1+z)\alpha + \rho}{(-1+z)\alpha + \rho + (-1+z)^2 z^{4+m}} \quad (16)$$

This equation clearly indicates the transfer function dependency on the loop latency m . In particular, the stability of the system, i.e. the locations of the closed loop poles, depends on m .

The compensated system, $\gamma = 1$ is now discussed.

Closed Loop Gain

The derivation of the closed loop transfer function is discussed. In this case, since

$\gamma = 1$, Equation (11) simplifies to Equation (17):



$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \alpha ((\theta in_{n-m-5} - \theta vco_{n-m-5}) - (\theta in_{n-m-6} - \theta vco_{n-m-6})) + \rho (\theta in_{n-m-6} - \theta vco_{n-m-6}) \quad (17)$$

which, recalling Equation (6) can be manipulated to yield Equation (18):

$$\theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} + \alpha (\theta vco_{n-5} - \theta vco_{n-6}) + \rho \theta vco_{n-6} = \alpha (\theta in_{n-m-5} - \theta in_{n-m-6}) + \rho \theta in_{n-m-6} \quad (18)$$

Taking the z-transform of Equation (18), we can obtain the following closed loop transfer function, expressed by Equation (19):

$$ClGain_1 = \frac{z^{-m} \alpha (z^{-5} + z^{-6}) + \rho z^{-6}}{1 - 2z^{-1} + z^{-2} + \alpha (z^{-5} + z^{-6}) + \rho z^{-6}} = z^{-m} \frac{(-1+z)\alpha + \rho}{(-1+z)^2 z^4 + \alpha (-1+z) + \rho} \quad (19)$$

Note that the term z^{-m} can be factored out indicating that, apart from a delay term, the closed loop transfer function is equivalent to Equation (20):

$$\frac{(-1+z)\alpha + \rho}{(-1+z)^2 z^4 + \alpha (-1+z) + \rho} \quad (20)$$

Consequently, the equation is not dependent on loop latency m, but is only dependent on the latency of the inner loop, which is essentially the latency of the loop filter 432 and the Vco accumulator 436. In other words, the circuit of Figure 4 is mathematically equivalent to the circuit of Figure 5, and all stability considerations should be concerned solely with Equation (20). In Figure 5, the output of adder 416 is connected directly to adder 404.

Open Loop Gain



The open loop transfer function can be computed directly from Equation (10) which, considering (5) and (6) can be rewritten as Equation (21):

$$5 \quad \phi vco_{n-m} - 2 \phi vco_{n-m-1} + \phi vco_{n-m-2} = \alpha (\theta in_{n-m-5} - \theta in_{n-m-6}) + \rho \theta in_{n-m-6} \quad (21)$$

The z transform, Equation (22), is obtained as follows:

$$10 \quad olGain_1 = \frac{\alpha (z^{-5} - z^{-6}) + \rho z^{-6}}{1 - 2z^{-1} + z^{-2}} = \frac{(-1+z) \alpha + \rho}{(-1+z)^2 z^4} \quad (22)$$

which differs from Equation (13) because the delay term z^{-m} (the loop latency) is now absent.

Steady state error

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The steady state errors to some common inputs, for example the steady state error to a phase step η , is given by Equation (23):

$$\lim_{z \rightarrow 1} \left((z-1) \left(\frac{\eta z}{z-1} \right) \left(1 - z^{-m} \frac{(-1+z) \alpha + \rho}{(-1+z) ((-1+z) z^4 + \alpha) + \rho} \right) \right) = 0 \quad (23)$$

20

The steady state error to a frequency step w_0 is given by Equation (24):

$$\lim_{z \rightarrow 1} \left((z-1) \left(\frac{w_0 z}{(z-1)^2} \right) \left(1 - z^{-m} \frac{(-1+z) \alpha + \rho}{(-1+z) ((-1+z) z^4 + \alpha) + \rho} \right) \right) = m w_0 \quad (24)$$

25 Equation (24) leads us to a limitation of the present scheme, as shown in Figure 4, that the steady state error due to a frequency step is non-zero and depends on the latency of the loop m . The reason becomes apparent when Equation (5) is rewritten for $\gamma = 1$



and uses the relationship between θ_{vco} and ϕ_{vco} shown in Equation (6) to obtain Equation (25):

$$\lambda_{n-m} = (\theta_{in_{n-m-2}} - \phi_{vco_{n-m-2}}) = (\theta_{in_{n-m-2}} - \theta_{vco_{n-2}}) \quad (25)$$

5

Since λ plays the role of system error signal, Equation (25) indicates that the feedback loop as shown in Figure 5 is aligning the current phase of the Vco to the phase of the input signal m clock cycles earlier. Equation (24) is then the result.

- 10 Although the scheme of Figure 4 is a solution to our stated problem in the presence of a phase step input, improvement can be achieved in practical cases where uncertainties in the frequency of the input signal make that assumption unnecessarily restrictive.

If we expand the term $1-CIGain$ in Equation (24), we obtain Equation (26):

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$$\frac{(-1+z)^2 z^4 + \alpha (-1+z) + \rho - z^{-m} ((-1+z) \alpha + \rho)}{(-1+z) ((-1+z) z^4 + \alpha) + \rho} = \frac{(-1+z)^2 z^4 + \alpha (-1+z) (1 - z^{-m}) + \rho (1 - z^{-m})}{(-1+z) ((-1+z) z^4 + \alpha) + \rho} \quad (26)$$

- It is noted that the numerator of Equation (26) has to go to zero faster than $(z-1)$ for the limit in Equation (24) to yield zero. For this to happen, the coefficient of the integral gain ρ in the numerator must be either 0 or contain a $(z-1)$ term. When either condition is met, the system should be able to track a frequency step with zero steady state error.
- 20

Next, two examples will be illustrated, starting from the topology of Figure 4, evolving to a configuration capable of processing uncertainties in the frequency of the input signal.

- 25 As illustrated in Figure 6, and rewriting Equation (10), Equation (27) is achieved.

$$\theta_{vco_n} - 2\theta_{vco_{n-1}} + \theta_{vco_{n-2}} = \alpha (\lambda_{n-m-3} - \lambda_{n-m-4}) + \rho \phi_{err_{n-m-5}} \quad (27)$$



Equation (27) reflects the fact that the integral path is being driven by ϕ_{err} (scheme A).

Next, the open loop gain is illustrated.

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Open Loop Gain

From Equation (27) we obtain Equation (27a) expressing λ in terms of ϕ_{err} :

$$\begin{aligned}
 10 \quad \theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \\
 \alpha(\phi_{err_{n-m-4}} - \psi_{n-m-4} - \phi_{err_{n-m-5}} + \psi_{n-m-5}) + \rho\phi_{err_{n-m-5}} = \\
 \alpha(\phi_{err_{n-m-4}} - \phi_{err_{n-m-5}}) + \\
 \rho\phi_{err_{n-m-5}} - \alpha(\phi vco_{n-m-5} - \theta vco_{n-m-5} - \phi vco_{n-m-6} + \theta vco_{n-m-6}) = \\
 15 \quad \alpha(\phi_{err_{n-m-4}} - \phi_{err_{n-m-5}}) + \rho\phi_{err_{n-m-5}} - \\
 \alpha(\theta vco_{n-5} - \theta vco_{n-m-5} - \theta vco_{n-6} + \theta vco_{n-m-6})
 \end{aligned} \tag{27a}$$

Equation (27a) yields the open loop transfer function in the z-domain, expressed as Equation (28):

$$\begin{aligned}
 20 \quad OlGain_z = z^{-m} \frac{\alpha(z^{-5} - z^{-6}) + \rho z^{-5}}{1 - 2z^{-1} + z^{-2} + \alpha(z^{-5} - z^{-6}) + \rho(z^{-5} - z^{-6})} = \\
 \frac{(-1+z)\alpha + \rho}{(-1+z)(-\alpha + z^m((-1+z)z^4 + \alpha))}
 \end{aligned} \tag{28}$$

Closed Loop Gain

Equation (27) in view of Equation (11) is transformed to Equation (29):

25

$$\begin{aligned}
 \theta vco_n - 2\theta vco_{n-1} + \theta vco_{n-2} = \\
 \alpha((\theta in_{n-5} - \phi vco_{n-m-5}) - (\theta in_{n-m-6} - \phi vco_{n-m-6})) + \\
 \rho(\theta in_{n-m-6} - \theta vco_{n-m-6}) = \\
 \alpha((\theta in_{n-m-5} - \theta vco_{n-5}) - (\theta in_{n-m-6} - \theta vco_{n-6})) + \rho(\theta in_{n-m-6} - \theta vco_{n-m-6})
 \end{aligned} \tag{29}$$



The closed loop transfer function in the z-domain is illustrated in Equation (30).

$$TlGain_{20} = \frac{z^{-m} \alpha (z^5 - z^{5+m} - \rho z^{-5})}{z^{4+m} - 2z^{3+m} + z^{2+m} + \alpha(z^5 - z^{5+m}) + \rho z^{2+m}} \quad (30)$$

- 5 If we compare Equation (30) to the one obtained for the topology of Figure 4, Equation (19), the term z^m can no longer be factored out. The consequence of this is that the loop latency will be only partially compensated, but the system will deal with the frequency uncertainties as discussed above.

10 Steady state error

The error gain is illustrated in Equation (31):

$$1 - ClGain_{20} = \frac{(-1+z) \{-z^{4+m} + z^{5+m} - \alpha + z^m \alpha\}}{z^{4+m} - 2z^{3+m} + z^{2+m} - z^m \alpha + z^{1+m} \alpha + \rho} \quad (31)$$

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Equation (31) is no longer dependent on ρ in the numerator. The steady state error in response to a phase step is defined by Equation (32):

$$\lim_{z \rightarrow 1} \left((z-1) \frac{\eta z}{z-1} \frac{(-1+z) \{-z^{4+m} + z^{5+m} - \alpha + z^m \alpha\}}{z^{4+m} - 2z^{3+m} + z^{2+m} - z^m \alpha + z^{1+m} \alpha + \rho} \right) = 0 \quad (32)$$

20

The steady state error to a frequency step ω_0 is Equation (33):

$$\lim_{z \rightarrow 1} \left((z-1) \frac{\omega_0 z}{(z-1)^2} \frac{(-1+z) \{-z^{4+m} + z^{5+m} - \alpha + z^m \alpha\}}{z^{4+m} - 2z^{3+m} + z^{2+m} - z^m \alpha + z^{1+m} \alpha + \rho} \right) = 0 \quad (33)$$



To gain an understanding of the loop operation, the loop operation in response to a frequency step is examined. Assume that the system has settled and the PLL is locked. At each clock cycle $phUpdate$ must increment ϕ_{vco} by $\Delta = \theta_{in_{n-1}} - \theta_{in_{n-2}}$.

- 5 However, because the loop is locked, $\phi_{err} = 0$, which implies Equation (33a):

$$\lambda_n = -\psi_{n-1} = -(\phi_{vco_{n-2}} - \theta_{vco_{n-m-2}}) = -m\Delta \quad (33a)$$

and the integral path accumulator settles to $(1 + \alpha m)\Delta$.

10

Finally, the steady state error to a frequency ramp of slope $\frac{\Delta}{2}$ is Equation (34):

$$\lim_{z \rightarrow 1} \left((z-1) \left(\frac{A z (z+1)}{2(z-1)^3} \right) \frac{(-1+z) \cdot (-z^{4+m} + z^{5+m} - \alpha + z^m \alpha)}{z^{4+m} - 2z^{5+m} + z^{6+m} - z^m \alpha + z^{1+m} \alpha + \rho} \right) = A \frac{(1+m\alpha)}{\rho} \quad (34)$$

15 Steady state error in presence of delay mismatches

Delay compensation approach includes removing the phase correction due to the vco phase update of time index $n-m-2$ as illustrated in Equation (4) and replacing it with the correction due to the phase update of time index $n-2$. As a consequence of delay mismatches, the cancellation may not be complete. For example, where m is the sum of two terms, a term associated with an analog delay and a term associated with a digital delay. The first term will, in general, pose a problem because it is not a well-defined multiple of the clock period due to its dependency on process and operating conditions.

25

In Figure 7, the delay mismatch has been introduced through the parameter δ and Equation (27) becomes Equation (35):



$$\begin{aligned}
& \theta vco_n - 2 \theta vco_{n-1} + \theta vco_{n-2} = \\
& \alpha \{ (\theta in_{n-m-5} - \theta vco_{n-m-5} - \psi vco_{n-m-5-\delta} + \theta vco_{n-m-5-\delta}) - (\theta in_{n-m-6} - \\
& \theta vco_{n-m-6} - \psi vco_{n-m-6-\delta} + \theta vco_{n-m-6+\delta}) \} + \rho \{ \theta in_{n-m-6} - \theta vco_{n-m-6} \}
\end{aligned} \quad (35)$$

The closed loop becomes Equation (36):

$$\begin{aligned}
& (z^{-m} (\alpha (z^{-5} - z^{-6}) + \rho z^{-6})) / (1 - 2z^{-1} + z^{-2} + \\
& \alpha ((z^{-m-5} + z^{-5-\delta} - z^{-m-5-\delta}) - (z^{-m-6} + z^{-6-\delta} - z^{-m-6-\delta})) + \rho z^{-m-6})
\end{aligned} \quad (36)$$

Equation (36) reduces to Equation (30) when $\delta = 0$. The error gain becomes Equation (37):

$$\begin{aligned}
& ((-1+z) (-z^{4+m+\delta} + z^{5+m+\delta} - \alpha + z^m \alpha)) / \\
& (z^{4+m+\delta} - 2z^{5+m+\delta} + z^{6+m+\delta} + \alpha - z \alpha - z^m \alpha + z^{1+m} \alpha - z^\delta \alpha + z^{1+\delta} \alpha + z^\delta \rho)
\end{aligned} \quad (37)$$

The steady state error in response to a phase step is Equation (38):

$$\begin{aligned}
& \lim_{z \rightarrow 1} \left((z-1) \frac{\eta z}{z-1} \right. \\
& \left. \frac{(-1+z) (-z^{4+m+\delta} + z^{5+m+\delta} - \alpha + z^m \alpha)}{z^{4+m+\delta} - 2z^{5+m+\delta} + z^{6+m+\delta} + \alpha - z \alpha - z^m \alpha + z^{1+m} \alpha - z^\delta \alpha + z^{1+\delta} \alpha + z^\delta \rho} \right) = 0
\end{aligned} \quad (38)$$

while the steady state error to a frequency step ω_o is Equation (39):

$$\begin{aligned}
& \lim_{z \rightarrow 1} \left((z-1) \frac{\omega_o z}{(z-1)^2} \right. \\
& \left. \frac{(-1+z) (-z^{4+m+\delta} + z^{5+m+\delta} - \alpha + z^m \alpha)}{z^{4+m+\delta} - 2z^{5+m+\delta} + z^{6+m+\delta} + \alpha - z \alpha - z^m \alpha + z^{1+m} \alpha - z^\delta \alpha + z^{1+\delta} \alpha + z^\delta \rho} \right) = 0
\end{aligned} \quad (39)$$

Finally, the steady state error to a frequency ramp of slope $\frac{4}{2}$ is Equation (40):



$$\lim_{z \rightarrow 1} \left((z-1) \left(\frac{A z (z+1)}{2 (z-1)^3} \right) \right) = \frac{(-1+z) (-z^{4+m+\delta} + z^{5+m+\delta} - \alpha + z^m \alpha)}{z^{4+m+\delta} - 2 z^{5+m+\delta} + z^{6+m+\delta} + \alpha - z \alpha - z^m \alpha + z^{1+m} \alpha - z^\delta \alpha + z^{1+\delta} \alpha + z^\delta \rho} = A \frac{(1+m\alpha)}{\rho} \quad (40)$$

Thus, we can conclude that the steady state system behavior is insensitive to delay compensation mismatches.

5

By forcing the delay compensated phase error term to zero, the scheme of Figure 4, in the presence of a frequency step, has a non zero steady state error. The concept of this second scheme lies in trying to compensate for this steady state error by feeding back an estimate of this frequency step. Such an estimate can be based on the content
10 of the frequency accumulator, f_{err} , with an appropriate scaling μ . A block diagram of such a system is illustrated in Figure 8 (scheme B).

In Figure 8, a multiplier 802 is added from the output of adder 412 to the summation circuit 404. To obtain an expression for the closed loop gain, Equations 41, 42 and 43
15 apply:

$$\psi_n = \phi vco_{n-1} - \theta vco_{n-1} = \phi vco_{n-1} - \phi vco_{n-m-1} \quad (41)$$

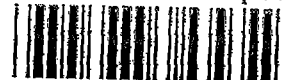
$$\lambda_n = \phi err_{n-1} + i_{n-1} - \psi_{n-1} \quad (42)$$

20

$$i_n - i_{n-1} = \mu (ferr_{n-1} - ferr_{n-2}) = \mu \rho \lambda_{n-3} \quad (43)$$

These equations, together with Equations (6) and (10), allow Equations (44), (45) and (46) in the z-domain:

25



$$\Psi = \Theta_{vco} (z^{-1} - z^{-n-1}) = \frac{\Theta_{vco}}{z^{-n}} (z^{-1} - z^{-n-1}) = \Theta_{vco} (z^{n-1} - z^{-1}) \quad (44)$$

$$\mathcal{L} = \Delta \mu \rho \frac{z^3}{1-z^{-1}} \quad (45)$$

$$\Theta_{vco} = \Delta z^{-n-4} \frac{\alpha(z-1) + \rho}{1-2z^{-1}+z^{-2}} \quad (46)$$

Combining these Equations (44), (45) and (46), we obtain Equation (47):

$$\begin{aligned} \Theta_{vco} &= z^{-n-4} \frac{\alpha(z-1) + \rho}{1-2z^{-1}+z^{-2}} (\Theta_{err} z^{-1} + \mathcal{L} z^{-1} - \Psi z^{-1}) = \\ &= z^{-2} z^{-n-4} \frac{\alpha(z-1) + \rho}{1-2z^{-1}+z^{-2}} (\Theta_{in} - \Theta_{vco}) + \\ &= \Theta_{vco} \mu \rho \frac{z^3}{1-z^{-1}} z^{-1} - z^{-n-4} \frac{\alpha(z-1) + \rho}{1-2z^{-1}+z^{-2}} \Theta_{vco} (z^{n-1} - z^{-1}) z^{-1} \end{aligned} \quad (47)$$

10

from which is obtained the closed loop gain in Equation (48):

$$ClGain_{21} = \frac{z^{-n} ((-1+z) \alpha + \rho)}{-\alpha + \rho + z ((-1+z)^2 z^3 + \alpha + \mu \rho - \mu z \rho)} \quad (48)$$

15 which reduces to (19) when $\mu = 0$.

Open loop gain

The open loop gain is shown in Equation 49.

20

$$OlGain_{21} = \frac{(z-1) \alpha + \rho}{z^n (-\alpha + \rho + z ((z-1)^2 z^3 - \mu \rho z + \alpha + \mu \rho)) - ((z-1) \alpha + \rho)} \quad (49)$$

Steady state error



The error gain is shown in Equation (50):

$$1 - \text{ClGain}_{21} = 1 - \frac{z^{-m}((-1+z)\alpha + \rho)}{-\alpha + \rho + z((-1+z)^2 z^3 + \alpha + \mu\rho - \mu z\rho)} \quad (50)$$

5

The steady state error in response to a phase step is shown in Equation (51):

$$\lim_{z \rightarrow 1} \left((z-1) \frac{\eta z}{z-1} \left(1 - \frac{z^{-m}((-1+z)\alpha + \rho)}{-\alpha + \rho + z((-1+z)^2 z^3 + \alpha + \mu\rho - \mu z\rho)} \right) \right) = 0 \quad (51)$$

10 while the steady state error to a frequency step is Equation (52):

$$\lim_{z \rightarrow 1} \left((z-1) \frac{\omega_o z}{(z-1)^2} \left(1 - \frac{z^{-m}((-1+z)\alpha + \rho)}{-\alpha + \rho + z((-1+z)^2 z^3 + \alpha + \mu\rho - \mu z\rho)} \right) \right) = \frac{(m-\mu)\omega_o}{(m-\mu)\omega_o} \quad (52)$$

which vanishes for $m \equiv \mu$.

15

The steady state error to a frequency ramp of slope $\frac{A}{2}$ is Equation (53):

$$\lim_{z \rightarrow 1} \left((z-1) \left(\frac{A z (z+1)}{2 (z-1)^3} \right) \left(1 - \frac{z^{-m}((-1+z)\alpha + \rho)}{-\alpha + \rho + z((-1+z)^2 z^3 + \alpha + \mu\rho - \mu z\rho)} \right) \right) = \frac{A(-2-2m\alpha+3m\rho+m^2\rho)}{2\rho} = \frac{A m (3+m)}{2} - \frac{A(1+m\alpha)}{\rho} \quad (53)$$

20 Figure 9 shows a modified version of the block diagram in Figure 1. Figure 9 illustrates a split latency m in two terms j and k , delay backs 902 and 904. The first one, delay back 904, represents the delay from the time the input signal is sampled to the time the corresponding phase error is computed and outputted from the phase detector. The second term k , delay back 902, represents the time that elapses between a Vco phase
25 update and the time the sampling position moves accordingly.



Figure 10 shows the corresponding delay compensated version (scheme A). In this case, the closed loop transfer function is given by Equation (19) with $m = j = k$ and that, as a consequence, the results derived for scheme A apply to Figure 12.

5

By inspection, we can obtain Equations (54), (55) and (56):

$$\theta vco_n = \psi vco_{n-k} \quad (54)$$

$$10 \quad \theta vco_n = \theta vco_{n-1} + phUpdate_{n-k-1} \quad (55)$$

$$\lambda_n = \phi err_{n-j-1} - \gamma \psi_{n-1} = (\theta in_{n-j-2} - \theta vco_{n-j-2}) - \gamma (\psi vco_{n-2} - \psi vco_{n-j-k-2}) = (\theta in_{n-j-2} - \gamma \psi vco_{n-2}) - \theta vco_{n-j-2} (1 - \gamma) \quad (56)$$

Equation (48) is similar to Equation (8) except for an index change. Consequently,

15 Equation (57) is illustrated:

$$\theta vco_n - 2 \theta vco_{n-1} + \theta vco_{n-2} = \alpha (\lambda_{n-k-3} - \lambda_{n-k-4}) + \rho \lambda_{n-k-4} \quad (57)$$

Using the new definition for λ given in Equation (49), we have Equation (58):

20

$$\begin{aligned} & \rho ((\theta in_{n-j-k-6} - \gamma \psi vco_{n-k-6}) - \theta vco_{n-j-k-6} (1 - \gamma)) = \\ & \alpha ((\theta in_{n-j-k-5} - \gamma \psi vco_{n-5}) - \theta vco_{n-j-k-5} (1 - \gamma)) - (\theta in_{n-j-k-6} - \gamma \psi vco_{n-6}) + \\ & \theta vco_{n-j-k-6} (1 - \gamma)) + \rho ((\theta in_{n-j-k-6} - \gamma \psi vco_{n-6}) - \theta vco_{n-j-k-6} (1 - \gamma)) = \\ & \alpha ((\theta in_{n-j-k-5} - \gamma \psi vco_{n-5}) - \theta vco_{n-j-k-5} (1 - \gamma)) - (\theta in_{n-j-k-6} - \gamma \psi vco_{n-6}) + \\ & \theta vco_{n-j-k-6} (1 - \gamma)) + \rho ((\theta in_{n-j-k-6} - \gamma \psi vco_{n-6}) - \theta vco_{n-j-k-6} (1 - \gamma)) \end{aligned} \quad (58)$$

25 For $\gamma = 0$, which is for the uncompensated case, Equation (51) reduces to Equation (59):



$$\begin{aligned} \theta vco_n - 2 \theta vco_{n-1} + \theta vco_{n-2} = \\ \alpha ((\theta in_{n-j-k-5} - \theta vco_{n-j-k-5}) - (\theta in_{n-j-k-6} - \theta vco_{n-j-k-6})) + \\ \rho (\theta in_{n-j-k-6} - \theta vco_{n-j-k-6}) \end{aligned} \quad (59)$$

which is identical to Equation (14) for $j + k = m$. For $\gamma = 0$, which is for the
 5 compensated case (scheme A), Equation (51) reduces to Equation (60):

$$\begin{aligned} \theta vco_n - 2 \theta vco_{n-1} + \theta vco_{n-2} = \\ \alpha ((\theta in_{n-j-k-5} - \theta vco_{n-5}) - (\theta in_{n-j-k-6} - \theta vco_{n-6})) + \rho (\theta in_{n-j-k-6} - \theta vco_{n-6}) \end{aligned} \quad (60)$$

Likewise, it can be demonstrated that the topology of Figure 11 is equivalent to the one
 10 presented in Figure 8 (scheme B).

The advantages associated with decoupling the PLL behavior from the loop
 latency can be summarized as follows. First, it is possible to introduce pipelining in the
 design of the sampling network (e.g. ADC) equalization circuits (e.g. finite impulse
 15 response filters) and decision directed phase detector of the read channel. This will
 ease the design of these critical components without degrading the performance of the
 PLL.

The present invention increases the system stability region. Because the
 20 stability region is significantly larger than in the uncompensated case, the tracking
 optimization is improved.

Figure 12 shows the stability region for an uncompensated system with $m = 10$.
 This contour plot has been generated by compiling, for a given α and ρ , the poles of
 25 Equation (16) and reporting the magnitude of the pole further from the origin. Clearly,
 those settings for which such value is ≤ 1 will correspond to stable operating points.
 For the purpose of this analysis, the ranges of the loop parameters ρ (x-axis) and α



(y-axis) have been adjusted to maximize the operating region, which is the region in which the system is stable.

Figures 13 and 14 show the stability region when using the topologies presented in Figures 6 and 7. When compared to Figure 12, it can be concluded that the stability region has increased by a factor of 20.

The dynamic characteristics of the system are greatly improved with respect to the uncompensated case. In particular, the residual steady state error in response to a frequency ramp is largely reduced because the system effective delay has been greatly reduced. Figures 15-17 show a comparison of the dynamics of the system to a phase step, a 1% frequency step and a frequency ramp, respectively. As in the previous case, the simulations were performed assuming $m = 30$.

The loop bandwidth is no longer limited by the loop latency and can be optimized on the basis of input noise and reference clock noise characteristics. In particular, in situations where the reference clock is noisy, the compensated topology presents a clear advantage, as illustrated in Figure 18. One can observe that the latency, by which the control action in response to noise on the reference clock is limited, is effectively canceled or significantly reduced by the compensated topology.



CLAIMS

1. A phase locked loop circuit for eliminating loop latency, comprising:
an input circuit for inputting a phase signal corresponding to phase changes and
5 an actual phase signal corresponding to actual phase sampling of a sampling clock;
a filter circuit for connecting said phase signal and said actual phase signal to a
phase update signal; and
a feedback circuit to eliminate said loop latency by feeding back to said filter-
circuit a portion of said phase update signal.

10



This Sheet Has Been Replaced

