THE PERFORMANCE OF A FAST CLOSED ORBIT FEEDBACK SYSTEM WITH COMBINED FAST AND SLOW CORRECTORS*

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Abstract

For NSLSII closed orbit feedback system, in order to limit the noise caused by digital step changes of the power supplies in the feedback system, the angular kick corresponding to the last bit of the power supplies for the fast correctors must be smaller than 3 nrad [1]. On the other hand, to carry out closed orbit alignment or orbit correction after a long term drift, we need strong correctors with 0.8 mrad kick strength [1]. In order to avoid the requirement of correctors with both large strength and very small minimum step size, we consider separate sets of slow correctors with large strength and fast correctors with smaller maximum strength. In order to avoid fast and slow feedback systems working in parallel, and avoid the possible interaction between two feedback systems, we consider the possibility of using only one fast feedback system with slow correctors periodically removing the DC components of the fast correctors so that the DC components in fast feedback system do not accumulate to reach saturation even after a large long term drift of the closed orbit motion. We report on simulation of the performance of such a combined system for NSLSII in this paper.

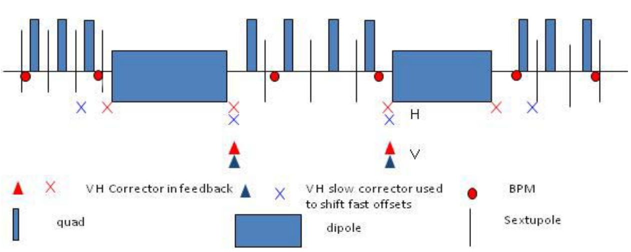

Figure 1: Layout of the orbit feedback system.

SYSTEM TRANSFER FUNCTION AND DC PERFORMANCE

In fig.1 we give the lattice positions of fast correctors and BPMs in the closed orbit feedback system and the slow correctors. In Fig. 2 we give schematics of the fast feedback system with slow correctors. R is the response matrix of the ring to fast correctors, R_B is the BPM part of R, R_s is the ring's response to slow correctors.

By singular value decomposition [2], $R_B = UW\tilde{V}$. The frequency transfer function of BPMs, fast correctors, the PID feedback circuit, and slow correctors are denoted by F_B, F_V, $-F_P$, and F_s, respectively, while the slow corrector strength (angular kick) is t_s. When the feedback loop is open, the BPM signals are given by:

$$y_{B0} = R_{BQ} Q y_{err} + R_{B\theta}\theta_{err} - y_{errB} + y_{Bnoise} \ .$$

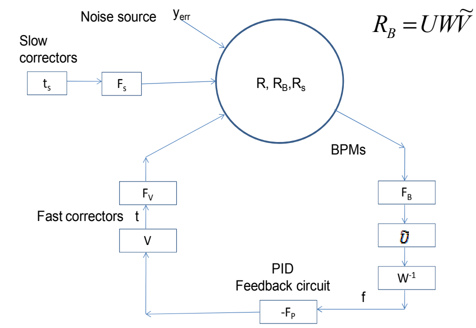

Figure 2: The fast feedback system with slow correctors.

Where y_{err} is the quads center motion, y_{errB} is the BPM motion, y_{Bnoise} is the BPM electronic noise, θ_{err} represent angular kicks from stray field errors, while R_{BQ} and $R_{B\theta}$ are the response matrices between BPM and quads and stray field errors sources, and Q is the quads KL value. From fig. 2 we see that when the feedback loop is closed, we have the BPMs signal:

$$y_B = R_B(-F_V V F_P W^{-1}\tilde{U} F_B y_B) + R_{BS} F_S t_S + y_{B0} \,,$$

where R_{BS} is the response matrix between BPMs and slow correctors. Now we assume F_B, F_V, $-F_P$ are the same for all channels, so they are just numbers, not matrices, and can be factored out and written as $F = F_B F_V F_P$.

Using [2], with $\tilde{V}V = \tilde{U}U = 1$, and $R_B V W^{-1}\tilde{U} = UW\tilde{V}VW^{-1}\tilde{U} = U\tilde{U}$, we get $(1+FU\tilde{U})y_B = R_{BS}F_S t_S + y_{B0}$. Since $U\tilde{U} \neq 1$ is a non-diagonal matrix not diagonal, instead of solving for y_B we solve for the feedback signal $f \equiv W^{-1}\tilde{U}F_B y_B$, with the result $(1+F)f = f_0 + W^{-1}\tilde{U}F_B R_{BS} F_S t_S$, where f_0 is the feedback signal when the feedback loop is open. Thus when the feedback loop is closed and the slow correctors are off, we have $f=f_0/(1+F)$, while the fast corrector strength is $t=-VF_P\ f_0/(1+F)$. After averaging over long time the DC components of t are used as the set points of slow correctors: $t_s=-V\ f_0G/(1+G)$, where $G=F(0)$ is the DC value of F. With a simple derivation we have the fast corrector strength when the slow correctors are on:

$$t = -\frac{F_P}{(1+F)} V\left[(1 - \frac{G}{1+G} F_B F_S) f_0 + F_B F_S W^{-1}\tilde{U}(R_{BS} - R_B) t_S \right].$$

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For DC, F_B, F_V =1, and F_P=G>>1, and (1-G/(1+G)) approaches 0 for large G, so the first term in the parenthesis is small. For the second term in the parenthesis, when the slow correctors are close to the corresponding fast correctors, R_{BS} is nearly equal to R_B, hence it is also very small. As result the fast correctors strength are largely reduced when the slow correctors are on. With these set points, after simple derivation, we get

$$y=\left(y_0-\frac{F}{(1+F)}RVW^{-1}\tilde{U}y_{B0}\right)+\left(R_S F_S t_S-\frac{F}{(1+F)}RVW^{-1}\tilde{U}R_{BS}F_S t_S\right).$$

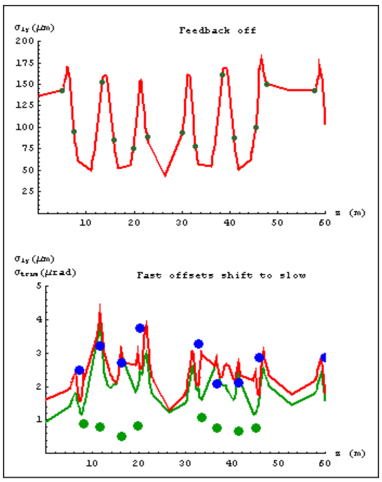

Figure 3: DC components of the residual orbit.

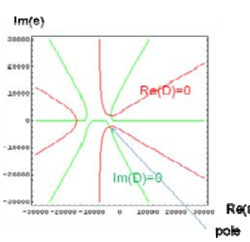

Figure 4: There are three poles on the complex plane of s, when τ =0.

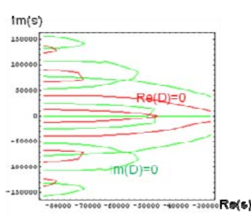

Figure 5: There are infinite number of poles on the complex plane of s, when τ =0.12ms

This is the residual orbit. The second terms in the two parentheses are the time dependent response of the fast feedback system to the noise and random motion of magnets and the response to the turning on of the slow correctors when their set points are sent in respectively. When we replace F by G in this expression we obtain the DC components of the residual orbit:

$$y=\left(y_0-\frac{G}{(1+G)}RVW^{-1}\tilde{U}y_{B0}\right)+\left(R_S t_S-\frac{G}{(1+G)}RVW^{-1}\tilde{U}R_{BS}t_S\right).$$

In the lower part of Fig.3 we plot the DC horizontal residual orbit. Here we assume the floor of the storage ring has a random long term drift with RMS value of 2 μ m, and there is a random floor vibration of 0.2 μ m around the storage and BPM noise of 0.2 μ m and the plot is the RMS values obtained with 400 random samples. The red line is the residual orbit when fast feedback is on but with fast correctors only, while the green line is with both fast and slow correctors on. The Blue dots are the strengths of the slow correctors while the green ones are for the fast correctors when both fast and slow correctors are on. If the slow correctors are turned off, then the fast corrector strengths would be represented by the blue dots. It is clear that the fast corrector strength is largely reduced when the slow correctors take on the set points of the fast correctors and the residual orbit is also slightly improved. If there is no further long term drift in the floor, and the set points of the fast correctors, after averaging over long time, are shifted to the slow correctors again, the residual strengths of the fast correctors would be further reduced. The residual strength of the fast feedback is not zero because of the 0.2 μ m floor vibration and BPM noise.

The result for vertical orbit is similar. The temporal response of the fast feedback system is determined by the factor F/(1+F)= F_B F_V F_P /(1+ F_B F_V F_P), while the response to the impulse of the slow correctors is determined by F_SF/(1+F). We have F_B $=\alpha_B/(s+\alpha_B)$,F_S $=\alpha_s/(s+\alpha_s)$, F_V= $e^{-s\tau}\alpha_V/(s+\alpha_V)$ respectively, with poles at $\alpha_B=2\pi\cdot 2kHz$, $\alpha_V=2\pi\cdot 1.5kHz$, $\alpha_s=2\pi\cdot 3Hz$, the delay in the power supplies including the effect of vacuum chamber next to the fast correctors is τ=0.2ms, and a trial feedback PID transfer function with pole position of at $\alpha_p=2\pi\times 2Hz$ and gain G=100: F_P $=G\alpha_p/(s+\alpha_p)$, where s is the Laplace transform variable of time t.

STEP FUNCTION RESPONSE

To obtain the response of the fast feedback system to a step function change of error field, we need to calculate the Laplace transform of

$$\frac{F}{(1+F)}=\frac{F_B F_P F_V}{1+F_B F_P F_V}=\frac{G\alpha_B\alpha_P\alpha_V e^{-s\tau}}{(s+\alpha_B)(s+\alpha_P)(s+\alpha_V)+G\alpha_B\alpha_P\alpha_V e^{-s\tau}}.$$

When delay is =0, the denominator is a third order polynomial, hence the three poles of the transfer function are easily obtained from the cubic equation

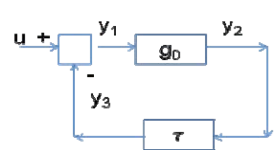

Figure 6: Equivalent circuit for response function.

$(s+\alpha_B)(s+\alpha_P)(s+\alpha_V)+G\alpha_B\alpha_P\alpha_V$ =0. With the three poles known, the inverse Laplace transform is easily calculated using Mellin's inverse integral by loop integral around the three poles, which is simply the residue of the poles. However, we realized that when $\tau\neq 0$, there are infinite number of poles. To show this we plot the zeros of the real part and the imaginary part of the denominator on the complex plane of variable s in Fig. 4. The crossing points of the zero line of the real part and imaginary part of the denominator D are the poles. There are 3 poles when τ =0. However there infinite number of poles when τ =0.12ms, as shown in Fig.5. In addition to this difficulty, our attempts to sum over the residue of these poles failed because the sum is found to be divergent. Hence we need a different method than Mellin's inverse integral. For this we write the response function as:

$$\overline{y_3} \equiv \frac{F}{1+F}\overline{u} = \overline{u}e^{-st}\overline{g_0}/(1+e^{-st}\overline{g_0}) \ ,$$

where $\overline{u}$ is the Laplace transform of step function representing the sudden turn on of a DC noise source, while $\overline{g_0}$ the transfer function F when τ =0, which is easily obtained by residue of the Mellin's integral with three poles $\alpha_B, \alpha_V, \alpha_P$.

The equivalent circuit diagram for y_3 is given in Fig. 6. It is clear from this we have $\overline{y_2} = \overline{g_0}\,\overline{y_1}$, $\overline{y_3} = e^{-s\tau}\overline{y_2}$, and $\overline{y_1} = \overline{u} - \overline{y_3}$. So we have

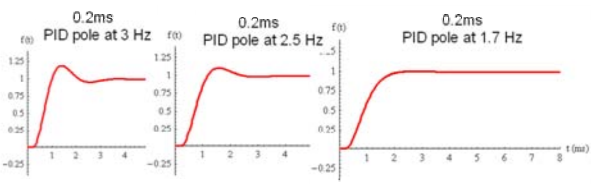

Figure 7: Step function response of fast feedback system.

$$\bar{y}_1 = \bar{u} - \bar{y}_3 = \bar{u} - \bar{y}_2 e^{-s\tau} = \bar{u} - \bar{g}_0\bar{y}_1 e^{-s\tau} \ .$$

This leads to $\bar{y}_1 = \bar{u}/(1+\bar{g}_0 e^{-s\tau})$. And hence indeed we have $\bar{y}_3 = \bar{g}_0 e^{-s\tau}\bar{u}/(1+\bar{g}_0 e^{-s\tau})$, as we expected. To obtain the Laplace transform of y_3, we write the Laplace transform of the 3 equations based Fig.6:

$$y_2(t) = \int_0^t g_0(t-t')y_1(t')dt',$$

$y_3(t) = y_2(t-\tau)$ and $y_1(t) = u(t) - y_3(t)$. These leads to integral equation

$$y_1(t) = u(t) - \int_0^{t-\tau} g_0(t-\tau-t')y_1(t')dt' .$$ This equation can easily be solved numerically by iteration: from time 0 to τ , y_3=0, so $y_1(t)$=u(t)=1, thus the integral in right can be calculated with t from τ to time 2 τ .

In turn, the result can be used the same way to calculated till t=3 τ . Continue this way, clearly we can calculate $y_1(t)$ to any time t. Then, y_3 is obtained by $y_3(t)$ = 1- $y_1(t)$. The step function response f(t)$\equiv y_3(t)$ for τ =0.2ms, with α_p=3Hz, 2.5Hz, and 1.7Hz is plotted in Fig. 7 showing when α_p=1.7Hz, the system reaches its DC level in shortest time (2ms). When α_p=3Hz there is an overshoot of nearly 20% and it takes 3ms to reach final value.

STEP FUNCTION RESPONSE OF FAST FEEDBACK SYSTEM COMBINED WITH SLOW CORRECTORS

With the Laplace transform of F/(1+F) obtained, we can combine this result with the performance of the system derived in page 2 for the residual orbit expressed by y to obtain its temporal behavior when the slow correctors are turned on for any specific random error of orbit drift. In Fig. 8 we plot the temporal response to the turn on of the slow correctors after there is a 2 m long term drift, random vibration and BPM noise are not included in this plot. At the center of the long straight section we can see the orbit position overshoots then return to its final value after about 10ms.

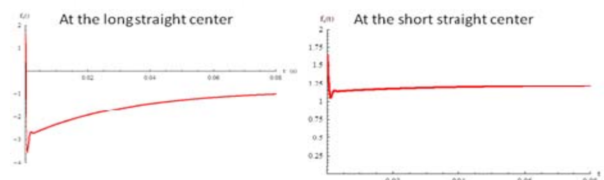

Figure 8: Temporal response to turn on of slow correctors at first long and short straight sections in one specific random example.

Notice that in this example we turn on the slow correctors to the set points of those fast correctors next to them without sending signals to the fast correctors at the same time. Clearly if we also send opposite signals to the fast correctors synchronously to remove the DC set points of the fast correctors, the temporal change of the orbit will be reduced nearly half to about 2 μ m from about 4 μ m. To eliminate this sudden change of orbit clearly we need to shift the set points of fast correctors to slow ones before the floor drift to far less than 0.2 μ m, the tolerance on the floor motion. Since our estimate based on the ATL law is that the floor motion around the ring within a day is about 4 μ m, we infer from ATL law that if the change of settings in the slow correctors is much more frequent than 3 minutes, the transient caused by this change will be much less than 0.2 μ m, and hence is negligible. We remark that the simulation in figure 3 is carried out assuming random floor motion, if we simulate the floor motion according to ATL law, the feedback system would be much more efficient in suppression of the orbit motion because the floor motion within short distance is correlated. In addition, the relevant length in applying ATL law is more close to beam line length than the whole ring. Hence our estimate is a conservative one.

REFERENCES

[1] NSLSII Preliminary Design Report (2007)
[2] W. Press, et.al., "Numerical Recipes" (1992)