Beam Feedback Systems

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Abstract. Outlines of bunch-by-bunch feedback systems for suppressing multibunch instabilities in electron/positron storage rings are shown. The design principles and functions of the feedback components are reviewed. The application of the feedback system as a tool to analyze instabilities using transient-domain techniques is also shown.

I INTRODUCTION

In recent high-beam-current, multibunch storage rings, especially particle factories such as B-factories, Φ-factories or synchrotron radiation facilities, beam instabilities are very strong because of many impedance sources such as RF cavities, bellows, radiation protection structures or special vacuum structures around interaction regions or insertion devices, and because of rather high beam currents. Moreover, the instability may have many unstable oscillation modes which are not easy to investigate or to avoid. Once instability occurs, it usually limits the maximum storable currents, reduces beam lifetimes, or enlarges the effective emittance in both the transverse and longitudinal planes, and spoils qualities of the beams such as luminosity or brilliance.

To overcome the instabilities, we must first reduce the sources of impedance. Higher-order modes (HOMs) coming from the RF cavities should be suppressed with a damped structure. Installing HOM absorbers near the high-Q impedance source will be effective in reducing the impedance. We should pay great attention to making the structures of vacuum components smooth. This is also effective in preventing breakdown due to heating, sparking, multipactoring and so on. The second method is to introduce a tune-splitting mechanism, which may be passive or active, to enhance the Landau damping effect. The third method is to employ beam feedback systems, and this is the main theme of this article.

In general, a beam feedback system consists of three main parts: a bunch oscillation detection system, a signal processing system that shifts the phase of the oscillation, cuts unnecessary signals and ensures synchronization between the feedback signals and the beam, and high-power amplifiers and kickers to kick the beam.

With a beam feedback system, we can first damp the coherent dipole instabilities whose growth rate exceeds radiation damping, Landau damping or head-tail damping. Also, we can damp injection oscillations much more quickly than with a radiation damping mechanism. Combining feedback systems and bunch oscillation memory systems, we can make a transient-domain analysis of the instability, which is a powerful tool for analyzing instabilities.

II SIMPLE EXAMPLE OF FEEDBACK SYSTEMS

As a simple example of a feedback system, we first consider an automobile cruise control system [1] (Fig. 1). If the driver keeps the gas pedal angle constant without

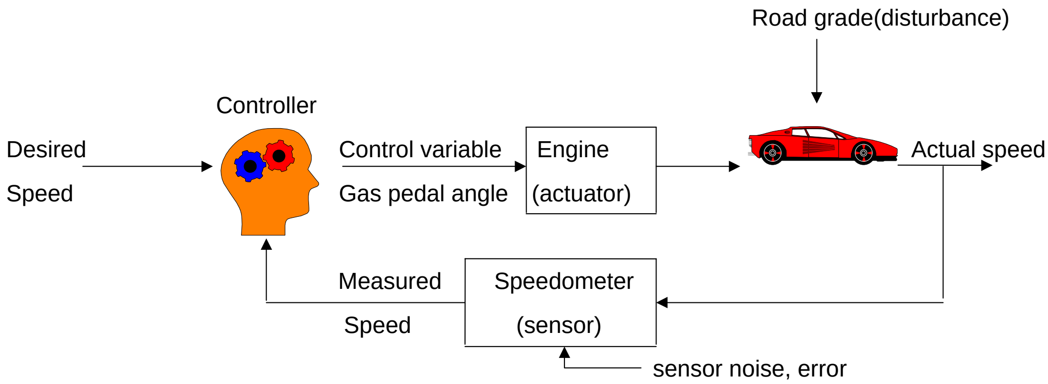


FIGURE 1. Component block diagram of automobile cruise control.

watching the speedometer, the speed will change with disturbances such as road grade or wind. This is called "open-loop control." Normally, the driver will watch the speedometer and tune the gas pedal angle to keep the speed constant. This is called "closed-loop control." Since he feeds back the speed information to the car, the effects of the disturbances will be greatly reduced. However, in some cases, for example if the driver is unskilled, the system can be unstable, and the speed may increase or decrease very rapidly, or introduce uncomfortable oscillations.

To see the effect of the closed-loop control system quantitatively, we need a set of mathematical relations among the variables of the system. For this example, we will ignore the dynamic response of the car and consider only the static behavior, though the dynamic response is also surely important in the system. We assume all the relations may be approximated as linear. For the vehicle, we measure speed on a flat road at 100 km/h and find that a unit change in the gas pedal angle causes a 10 km/h change in speed. For a grade change of 1%, the speed is reduced by 10 km/h. With these relations, we can draw the block diagram shown in Fig. 2(a).

In the open-loop case, Fig. 2(b), the output speed is given by

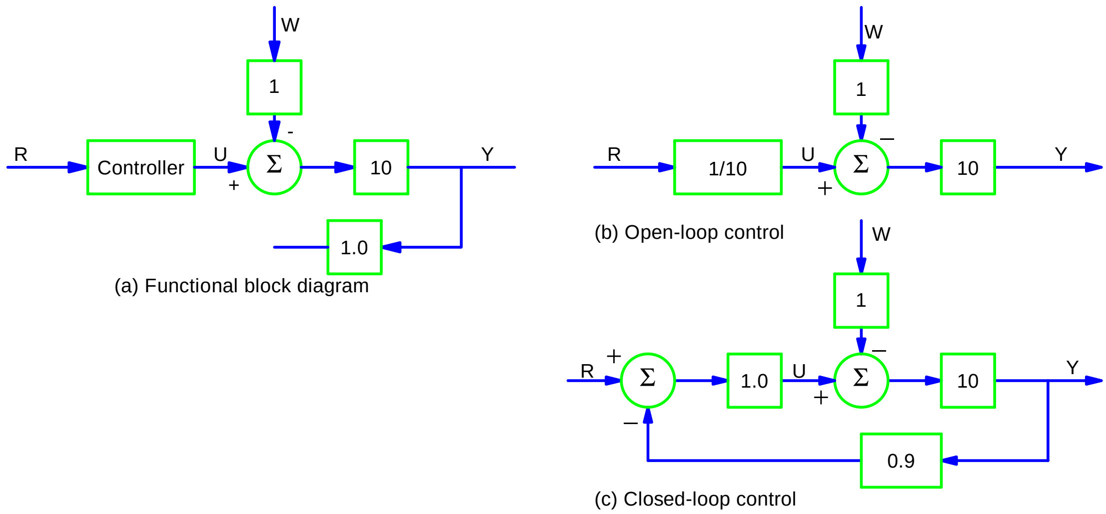


FIGURE 2. Block diagram of cruise control.

$$\begin{aligned} Y_{ol} &= 10(U - W) \\ &= R - 10W. \end{aligned}$$

If $W = 0$, on a flat road, and $R = 100$, then the speed will be 100 km/h and there are no errors. However, if $W = 1$, on a 1% grade, then the speed will be 90 km/h, and we have 10 km/h (10%) error. Let us compare this with the feedback scheme shown in Fig. 2(c). In this case the output speed is

$$\begin{aligned} Y_{cl} &= 10(U - W), \\ U &= R - 0.9Y_{cl}, \\ Y_{cl} &= R - W. \end{aligned}$$

In this case, if the input set-point R equals 100 km/h and the grade W is 1%, the speed will be 99 km/h, only 1 km/h (1%) error— the feedback works to reduce the error by a factor of 10.

We will not go further into the details of feedback theory [1] here, though a precise understanding of it is surely important. We note here two important points to consider. The first is the *positive* feedback situation. In the above example, we added the feedback speed information *negatively* to the controller. If we add the information *positively*, the error is not reduced but amplified. In this situation, it is in general not easy to control the system without a self-limiting (none-linear) element in the system. A frequency oscillator is an example of this application. The second point is feedback gain. Increasing the feedback gain, reduces the steady-state errors, and it provides faster response of the system with the gain. However, it also reduces the stability of the feedback system under realistic conditions. The system can easily begin to oscillate or go out of control. Recall the screeching of a sound system (usually called howling) when one raises the gain of the power amplifier too much. We must keep a gain margin and a phase margin in the feedback system to avoid unstable situations.

III BEAM FEEDBACK SCHEME

Figure 3 shows the phase-space view of a bunch-by-bunch feedback scheme. In

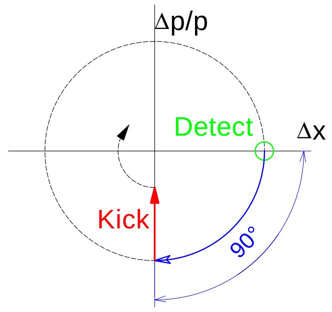


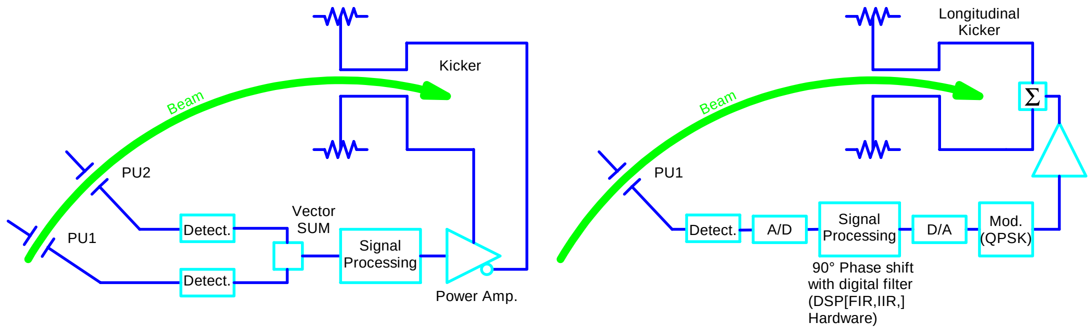
FIGURE 3. Phase-space view of the bunch-by-bunch feedback scheme.

the beam feedback system, we first detect the bunch position, then make a 90° phase shift to the beam, rejecting the unnecessary static component in the signal processing unit. Here the static component in the feedback loop may be the residual closed-orbit distortion (COD) for a transverse plane, or the change in equilibrium phase in the longitudinal plane. Next, we wait for the bunch re-arrival at the kicker section and kick the bunch to change the angular divergence in the transverse plane, or momentum in the longitudinal plane, by the feedback kicker.

Suppose our ring contains equally spaced bunches. The feedback system must be able to respond to an arbitrary pattern of oscillations of the bunches. If all bunches require a kick of the same amplitude and sign, this is the same as a DC correction. In reality, the minimum frequency is not 0 Hz but the base-band betatron frequency or synchrotron frequency, depending on the feedback plane. The highest possible frequency we will need comes from the case where all the subsequent bunches have equal magnitude with completely opposite sign. Thus, the upper end of the bandwidth must exceed half the bunch frequency, $f_b/2$. The center frequency of this bandwidth is somewhat arbitrary: when there are no bunches passing through the kicker, it does not matter what kick is applied.

Figure 4(a) shows a typical transverse feedback diagram. We detect the bunch positions by two beam position monitors (BPMs), which are arranged so that the betatron phase advance is around 90°. The two signals are summed vectorially to make a suitable phase shift to the feedback kicker, around 90°. With the signal-processing part we reject the DC component and adjust the delay, around one revolution period minus the signal delay in the circuit and cables.

Figure 4(b) shows a typical longitudinal bunch-by-bunch feedback system. We detect the longitudinal position of a bunch, which is the phase of the bunch. Since



(a) Typical transverse beam feedback system (b) Typical longitudinal bunch-by-bunch feedback system

FIGURE 4. Typical bunch-by-bunch feedback diagram.

the synchrotron frequency is much lower than the revolution frequency, we must make a 90° phase shift by using a digital filter. As in the transverse case, we also provide a one-turn delay here. In some cases, if we need to use a higher-frequency kicker, we need a modulation circuit before the amplifiers.

IV FEEDBACK SYSTEM HARDWARE

A Pickup electrodes

In the design of pickup electrodes for bunch feedback systems, we have several requirements, as follows:

- Vacuum safety structure.
  All the components should be strong enough to withstand typical mechanical shocks, such as welding, vacuum baking, small impact or cable related shocks. Additionally, it should be tough enough to withstand the extracted beam power or high peak-voltage. Trapped modes in the structure should be damped as much as possible.

- Sufficient, but not excessive output power.
  Though we need enough output from the beam for a good signal-to-noise ratio, unnecessary or excessive power is troublesome because it can burn out attenuators, RF components or cables.

- Wideband frequency response and clear impulse response.
  For the separation of signals from two successive bunches, we need wideband and clear response.

Figure 5 shows a button-type electrode used in the KEKB feedback systems and its simplified equivalent circuit. It forms a high-pass filter with a time constant of τ =1/CR. Now, there are good simulation codes such as HFSS [2] and MAFIA [3] for calculating the RF characteristics with good accuracy without the need to manufactured test specimens.

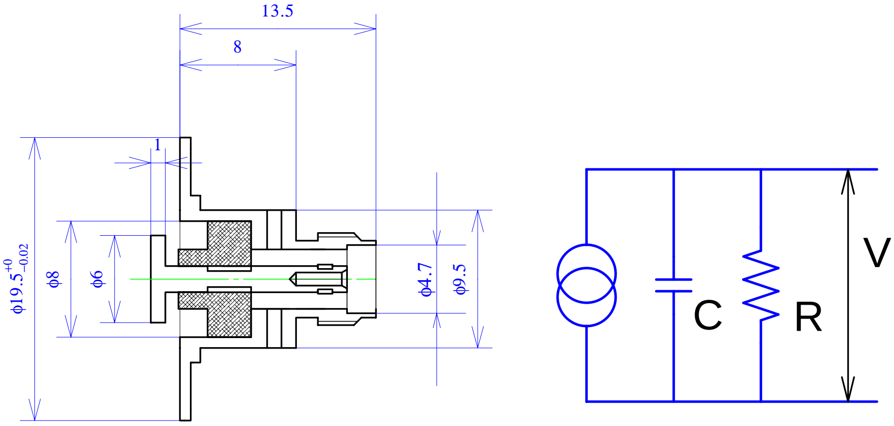


FIGURE 5. KEKB button electrode for feedback systems and its equivalent circuit.

B Bunch position detection electronics

The requirements for bunch position detection electronics are as follows:

- Quick output.
  Since we must determine the proper feedback kick voltage before the bunch arrives at the kicker section, the processing time of the detection circuit should be much shorter than the revolution period. Consequently, the detection circuit should be a simple hardware system without any software corrections.

- Wideband response.
  To distinguish the signals from two adjacent bunches, the circuit should have a high enough bandwidth. We must remember to check the bandwidth and the step response of the RF components of the circuit.

- Sufficient signal-to-noise ratio.
  The signal level and the noise-figure of the amplifiers in the circuit should be well tuned.

To achieve these features, we must accept some undesirable features such as poor linearity or saturation at high amplitude. The absolute positions and gains are usually not important. And we can accept some bunch-current dependence of the output of the feedback if the dependence is not severe.

Figure 6 shows a longitudinal bunch position detection circuit. It consists of a band-pass filter (BPF) to make a sinusoidal burst-like signal from the BPM signal, a double-balanced mixer (DBM) that works as the multiplier between the two arbitrary inputs, a low pass filter (LPF), a frequency multiplier and a phase shifter.

We first extract the $n\omega_{RF}$ component, which is the detection frequency, from the beam signal with the BPF. If the beam is oscillating longitudinally with a

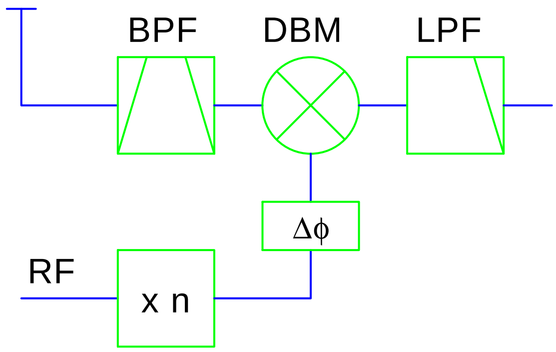


FIGURE 6. Longitudinal bunch-by-bunch position detection circuit.

synchrotron frequency of ω_s and an amplitude of Φ, the signal after the BPF will be expressed as

$$I_b \cos(n\omega_{RF}t + \Phi \sin \omega_s t).$$

By multiplying the signal with the n-th hermonic of RF synchronized signal 90° shifted from the beam phase by the phase shifter ($\sin(n\omega_{RF}t)$) using the DBM, we can split the signal into lower and higher frequency components as

$$\begin{aligned} & I_b \cos(n\omega_{RF}t + \Phi \sin \omega_s t) \times \sin(n\omega_{RF}t) \\ = & \frac{1}{2} I_b \left(\sin(2n\omega_{RF}t + \Phi \sin \omega_s t) - \sin(\Phi \sin \omega_s t) \right). \end{aligned}$$

By rejecting the higher frequency component with the LPF, and assuming the amplitude of the oscillation to be small enough, the final output is proportional to the longitudinal displacement of the bunch as

$$\propto I_b \sin(\Phi \sin \omega_s t) \sim \Phi \sin \omega_s t.$$

Note that the sensitivity is roughly proportional to the detection frequency and the output is proportional to the bunch current. The total step response of the system is in many cases dominated by the bandwidth and the response function of the first bandpass filter since the bandwidths and step responses of other components have sufficient margins compared to the required specifications.

A transverse bunch detection circuit is shown in Fig. 7. The difference between the beam-induced signals of the two facing electrodes, which is proportional to the transverse displacement of a bunch, is made by using a hybrid circuit. Contrary to the longitudinal case we need a phase-independent component, so we multiply the $n\omega_{RF}$ signal with nf_{RF} in-phase using the DBM:

$$\begin{aligned} & I_b \cos(n\omega_{RF}t + \Phi \sin \omega_s t) \times \cos(n\omega_{RF}t) \\ = & \frac{1}{2} I_b \left(\cos(2n\omega_{RF}t + \Phi \sin \omega_s t) + \cos(\Phi \sin \omega_s t) \right) \\ \rightarrow & \frac{1}{2} I_b. \end{aligned}$$

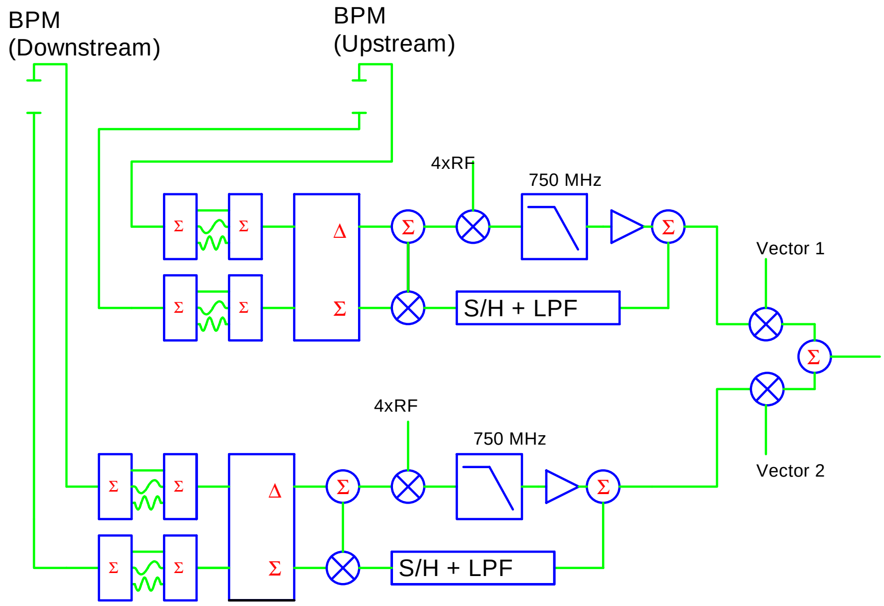


FIGURE 7. Transverse bunch by bunch position detection circuit.

C Signal processing

Since the speed of signals in the cables, detection electronics or amplifiers is much slower than that of beam, and as we want to concentrate the feedback components in one section of the ring, we need to wait for the beam to come back to the feedback section after one turn. In smaller rings, typically of circumference less than 300 m, we can use cable delay. For large rings, such as the KEKB, with a circumference of 3 km, the use of a cable delay of 10 μs is hopeless. In that case we can use a digital delay which converts the analog signal to a digital signal with a very fast ADC, records the digital signal in memory, re-reads it after some time and re-converts it to an analog signal with a very fast DAC.

The DC component in the feedback signal, which will be the residual COD in the transverse feedback, the change in the equilibrium phase or beam-loading effect in the longitudinal feedback, should be rejected as much as possible, because the feedback system tries to correct those errors, in vain. The DC signal wastes very "expensive" feedback power and causes undesirable rapid saturation at the ADC or the power amplifier section. To reject the DC component in the loop, there are several useful methods. For a small ring using long cable delay for one-turn delay, we can form a notch filter to reject the nf_{rev} component combining the cable delay and analog subtracter. Note, however, that this type of cable, with a very long cable and tuning capacitors, is sensitive to temperature changes of the surroundings so careful handling is necessary. For a large ring with digital delay, we need to take great care to reject the offset before the ADC, because the dynamic range of a fast ADC is very limited, in many cases only 8-bit. For automatic correction, we can add offset suppressor feedback loops with very slow local feedback loops before the ADC. Within the dynamic range of the ADC, we can form a DC rejection filter in the digital signal processing logic.

There are two approaches to making a digital filter. The first is to use a software-based system using many digital signal processors (DSPs) [4]. Taking advantage of the great progress in digital circuit technology, using DSPs of the highest execution speed enables us to design a complicated floating-point digital filter with considerable speed. The strong points of a software-based signal process system are as follows:

- Completely programmable. It is easy to change the type of filter by replacing the DSP code.
- Good filtering characteristics. We can form multi-tap filters easily with good accuracy.
- Flexibility. It is applicable to both small rings and large rings by changing the number of DSPs on boards.

However, it also has such weak points as:

- Large-scale and complicated design. As the speed of a single DSP is much slower than the given time limit of a calculation, we require many parallel processing cards. For a fixed number of coefficients in the digital filter, the number of DSP cards increases with the number of bunches times the synchrotron frequency. The difficulty of tuning and maintenance correspondingly increases.
- It is in general necessary to use the down-sampling technique to avoid this complexity [4]. In this case we cannot employ the same system as for the transverse feedback.

Because of its great flexibility and established usability, many rings such as ALS (LBNL), PEP-II, SPEAR (SLAC), PLS (Pohang), DAΦNE (Frascatti) and BESSY-II (BESSY) are using the same system for longitudinal feedback. In PEP-II, 40 DSP's are used for one ring, which provide an aggregate multiple-accumulation rate of 1.6×10^9 operations/s.

The second approach, which we have employed for the KEKB rings, is to make the simplest digital filter that satisfies our minimum requirements with fast hardware logic circuits without the down-sampling technique [5]. The structure of the filter is the hardware two-tap FIR proposed F. Pedersen [6]. As shown in Fig. 8, it has only two taps, A(+1) at $-90°$ and B(-1), at $-270°$ of the oscillation. The frequency response has very wide peaks at f_s, $3 \times f_s$, $\cdots$ and has zeros at 0 (DC), $2 \times f_s$, $\cdots$, where f_s is the synchrotron frequency in the longitudinal case. Strong points of the hardware two-tap FIR filter are:

- Simple and applicable to both longitudinal and transverse feedback systems. The phase shift and time delay are tunable by selection of the tap positions while preserving the time differences between the two taps.

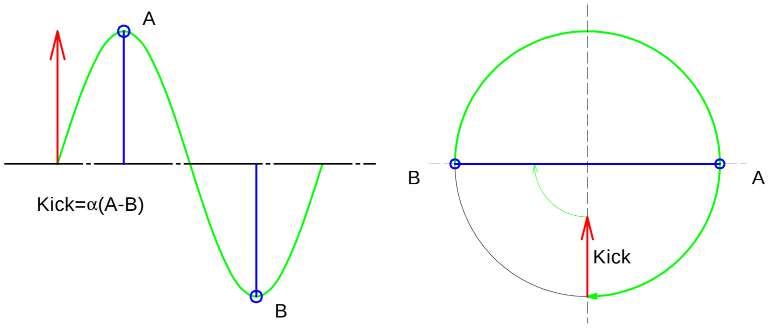


FIGURE 8. Tap positions for the 2-tap FIR filter.

With this simplification, the function of the filter is fairly limited and the following weak points arise:

- Very limited flexibility. To reduce the complexity, the structure of the filter needs to be strongly geared towards a particular ring. Application to other rings is in general very difficult.
- No sharp-filtering effect around the center frequency. This is, in practice, not a serious problem. Experiments with beams show that the measured S/N of the detection signal is good enough, typically better than 40 dB.
- Complication of the high-frequency digital circuits around the ADC and the DAC. At KEKB, these difficulties have been overcome by developing custom LSIs to multiplex and demultiplex the fast digital data.

Figures 9 and 10 show a block diagram and a photo of the two-tap FIR filter.

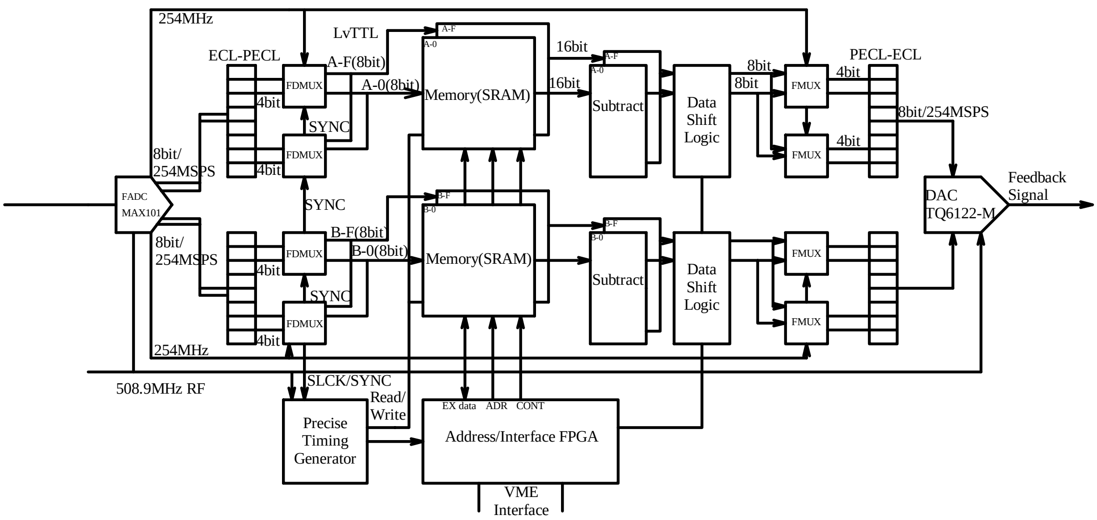


FIGURE 9. Block diagram of the 2-tap FIR filter board developed at KEK.

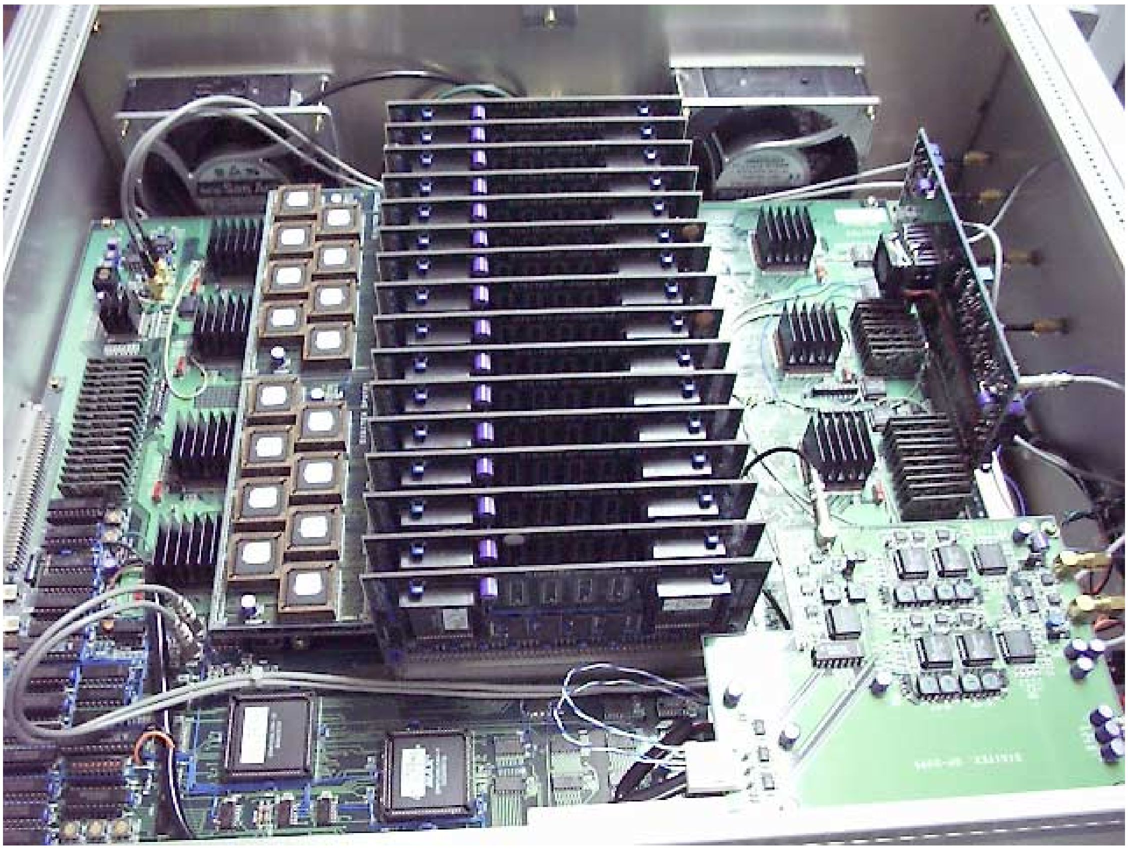


FIGURE 10. Photo of the 2-tap FIR filter board for KEKB.

D Kickers and power amplifiers

We first show rough formulae for estimating the necessary feedback voltage for a desired damping time. For the longitudinal case, to damp a synchrotron oscillation of maximum energy deviation ΔE with a feedback damping time of τ_ϵ, we need

$$V_{feedback} = 2\frac{1}{\tau_\epsilon}T_0(\Delta E/e)$$

where T_0 is the revolution time. For example, assuming E=3.5 GeV, $(\Delta E/e)/E = 0.1\%$, $T_0 = 10$ μs and $\tau_\epsilon = 10$ ms, we need 7 kV/turn, which is a fairly tough value to achieve.

For the transverse case, we have

$$V_{feedback} = 2\frac{1}{\tau_x}T_0(E/e)\frac{1}{\sqrt{\beta_m\beta_k}}x_{max}$$

where τ_x is the feedback damping time, x_{max} is the maximum amplitude, β_m and β_k are the betatron functions at the monitor and the kicker, respectively, assuming the betatron phase advance between the kicker and the monitor is 90°. Again, to damp x_{max} = 1 mm oscillation with τ_x=1 ms in the case of E =3.5 GeV, $\beta_m = \beta_k$ = 10 m and T_0 = 10μs, we need $V_{feedback}$=7 kV/turn. Though it depends on the mode of oscillation, this kick voltage is not an unreasonable value.

To act as wideband kickers, stripline-type kickers are widely used. The structure(Fig. 11) consists of a rod or a plate as inner conductor and the wall as outer conductor. The characteristic impedance of the kicker is normally designed to be

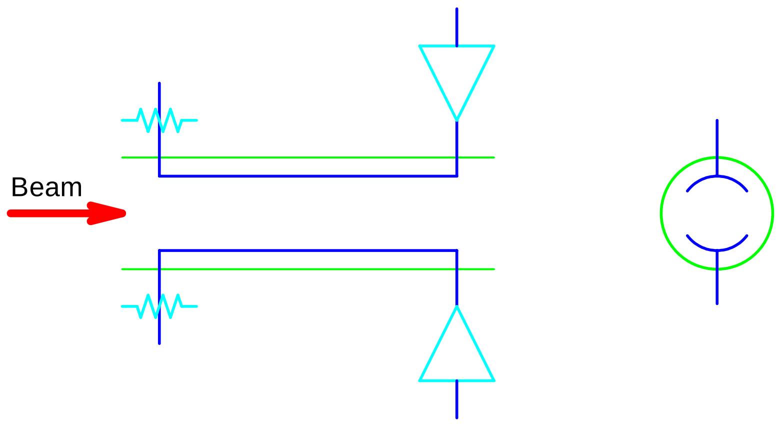


FIGURE 11. Sketch of a strip-line type kicker.

50 Ω. We know from the Panofsky-Wenzel theorem [7] that we need a change in the longitudinal electric field to kick the beam, even in the transverse direction. Therefore, the deflecting power must be supplied from the downstream port. If we supply the power in phase, we can kick the beam longitudinally. If the power is supplied with opposite phase, the beam will be deflected transversely. The shunt impedance of the kicker is calculated based on the direction of the kick. The longitudinal shunt impedance $R_\parallel$ is given [8] by

$$R_\parallel T^2 = 2 Z_L g_\parallel^2 \sin^2 k\ell,$$

where T^2 is the transit-time factor, Z_L is the characteristic impedance of the kicker, $g_\parallel$ is the longitudinal geometric factor, k is the wave number, and ℓ is the length of the stripline. The transverse shunt impedance $R_\perp$ is given by

$$R_\perp T^2 = 2 Z_L \left(g_\perp \frac{2\ell}{h} \frac{\sin k\ell}{k\ell} \right)^2 ,$$

where $g_\perp$ is a transverse geometric factor and h is the distance between the facing electrodes. From the formulas above, we know that:

- The longitudinal shunt impedance is very low, around 100 Ω. Since the shunt impedance is a periodic function of the frequency, we can use higher-frequency components which helps in the design of wideband power amplifiers.

- The transverse shunt impedance is proportional to $\sin^2(k\ell)/(k\ell)^2$ so we can use only the base-band region. A longer stripline has larger shunt impedance but the bandwidth decreases rapidly. Typical shunt impedance is around a few kΩ.

For longitudinal kickers, ALS and PEP-II have adopted a series-drift-tube type kicker [9], in principle a stripline kicker but combined with two electrodes to make good use of feedback power. The measured shunt impedance for one unit is about 320 Ω. At DAΦNE, an over-coupled cavity (namely very wideband) has been developed for the longitudinal kicker [10]. A quality factor less than 5 is achieved around the center frequency of 1 GHz with a shunt impedance greater than 600 Ω. This kind of longitudinal kicker has been adopted at DAΦNE, KEKB-LER, SPEAR, PLS and BESSY-II. Since this kind of kicker has no directivity, we need wideband circulators to protect the amplifiers from the beam-induced power.

The maximum power required for the feedback system is easily evaluated as

$$P_{MAX} = V_{MAX}^2/2R_{sh}$$

where V_{MAX} is the maximum feedback voltage. Actually, however, the maximum power of available wideband amplifiers is limited for technical and the economic reasons. The cost of power for a wideband amplifier is in general extremely high. We usually restrain ourselves to selecting smaller and cheaper amplifiers.

If the power of the amplifiers is sufficient, not saturating at the needed feedback voltage, the oscillation should be damped exponentially, with the designed feedback damping time. However, when the amplifiers are not powerful enough to supply the necessary power, the oscillation will be damped linearly, not exponentially. We call this regime "bang-bang damping." The effective damping time in this region is much longer than the feedback damping time. In some cases which the growth rate is huge, the feedback system fails to damp the oscillation after the oscillation exceeds some threshold amplitude (cannot re-capture the oscillation).

V TRANSIENT-DOMAIN ANALYSIS

The transient behavior of the beam just after closing or opening of the feedback loop reveals many important characteristics of the coupled-bunch motions as well as the performance of the feedback systems. This powerful method of analyzing instabilities is known as transient-domain analysis [4,11]. By preparing a large-scale memory board that can accumulate bunch positions for every bunch with many turns, we can record and analyze the time evolution of oscillation for every bunch at the transient of feedback on/off. Figure 12(a) shows the time evolution of the betatron-oscillation components for each bunch. By taking the Fourier transform of the bunch amplitude over time, taking into account the betatron phase advances, we calculate the evolution of the modes of the instability over time, as shown in Fig. 12(b).

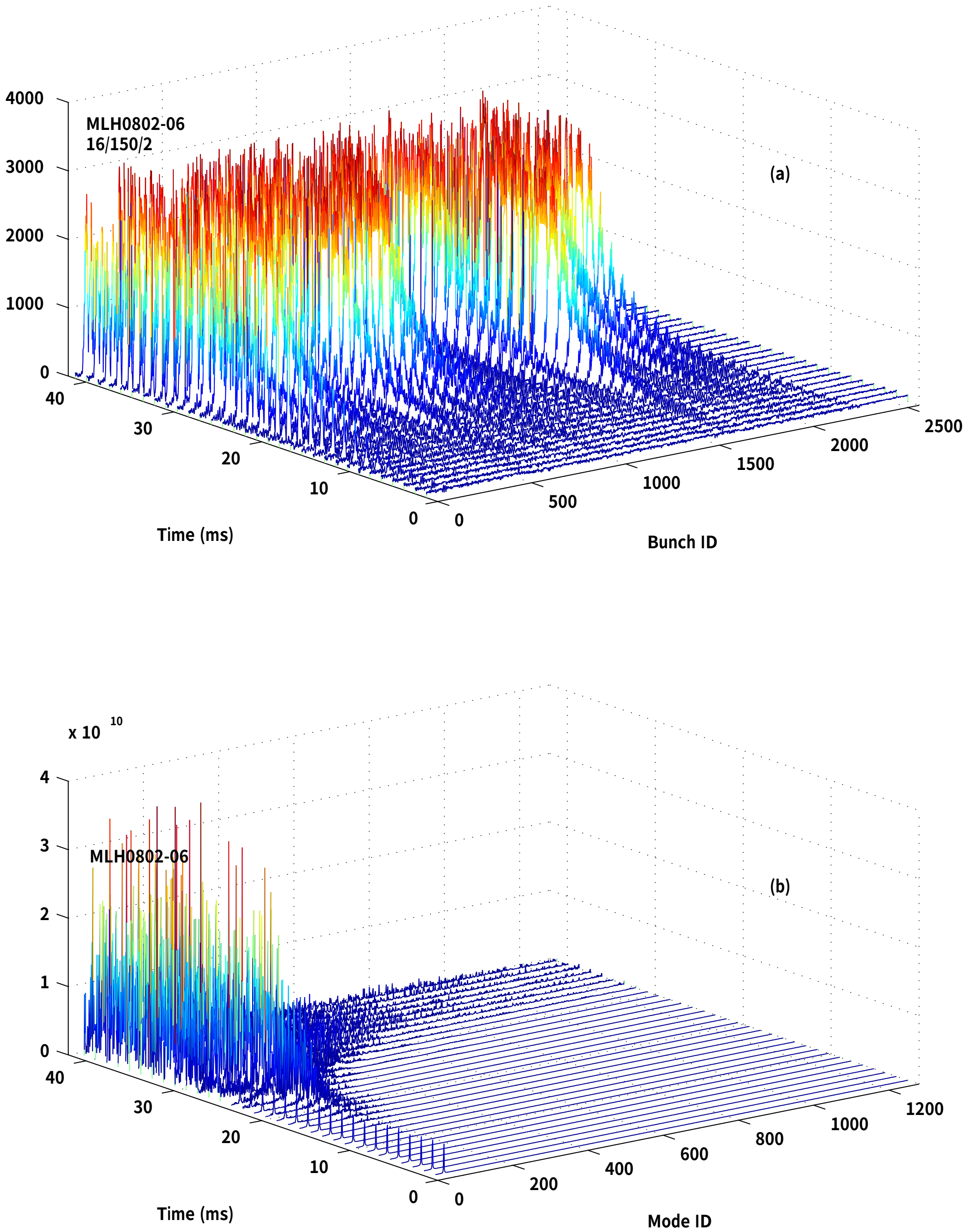


FIGURE 12. Example of a growing transient due to horizontal instability in the KEKB LER. (a) Time evolution of the betatron amplitude for each bunch; (b) Time evolution of the modes of the instability.

Comparing this method with the ordinary method of analyzing instabilities, we can get the full spectrum of bunch motion in a single transient. And the motion can be studied in the small-oscillation situation where the oscillation may be regarded as linear.

SUMMARY

We have seen the outline of the bunch-by-bunch feedback systems and the application of the feedback systems to analyze the instabilities. However, these complicated systems are also fairly expensive and require skills to operate them and keep them in good condition. You may ask, "Is such a costly and complicated system surely necessary?" I believe, that making all possible efforts to reduce impedance sources in the ring is surely valuable, even if the work is costly and boring. The power of bunch feedback systems is limited and they can damp only the dipole oscillation within some limited amplitude. However, we must also consider the total cost performance and the possibility of unknown broad-band impedance sources. With a beam feedback system we can suppress the instabilities coming from uncertain or unknown broad-band impedance sources. Also we can diagnose the impedance source by transient-domain analysis, helping to locate the source. Beam feedback systems can greatly contribute to the high quality operation of a ring.

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