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A new IQ detection method for LLRF

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ABSTRACT

Digital LLRF technology has been widely used in new generation particle accelerators. IF quadrature sampling is a common method for amplitude and phase detection. Many strategies, which obey the same rule of $f_{\mathrm{sample}}=(M/N)f_{\mathrm{IF}}$ (M/N is a rational number), have been proposed to reduce the effects of spectrum aliasing. However, we found that M/N does not need to be a rational number according to Shannon's theorem. Therefore, we propose a new IQ detection method in this paper. This method is based on a special IIR filter which is derived from an RLC circuit. The unique characteristic of the method is that the value of f_{IF} is independent of the value of f_{sample}. We have set up an experimental platform to verify our method. A 122.88 MHz sampling clock is used to sample a 3 MHz IF signal. The DDS and PI control techniques are used to realize the closed-loop control. Results show that the stability of the system is within $\pm$ 0.05% (peak to peak) for the amplitude, and with $\pm 0.03^{\circ}$ (peak to peak) for the phase in 5 h.

1. Introduction

The amplitude and phase stability of the RF field directly affects the quality of the beam in accelerators. With increasing requirements for higher quality beams, high precision feedback control is desperately demanded by LLRF systems. For example, International Linear Collider (ILC [1]) and X-Ray Free Electron Laser (XFEL [2]) require the RF amplitude stability of 0.07% and 0.01%, phase stability of 0.24° and 0.01°, respectively.

With the development of microelectronic technology, especially the high performance Field Programmable Gate Array (FPGA), many concepts and techniques have been introduced in the commercial and military fields. The 4G terminal and the software defined Radio are proposed in such a background. These two techniques and the digital LLRF have a similar structure, while the algorithm is the core function. An overview of a typical software defined Radio system is shown in Fig. 1. All these new techniques have set up a solid foundation for the development of high precision feedback control in RF control systems for modern accelerators.

High accuracy IQ detection is the basic for the high precision RF control. At present, the IQ [3] and non-IQ [4] are the best rounded sampling method in the main labs such as DESY, KEK and Fermilab [5]. This paper presents a new IQ detection method. Compared with the traditional method, the proposed method is based on a narrow band IIR filter with advantages in short latency and high flexibility.

2. The new method for IQ detection

A rotation vector algorithm has been used in this IQ detection method. The continuous-time IF signal $f(t)$ can be represented by a discrete-time signal consisting of a sequence of samples $f(n)=A\cos(\Omega_0 T\cdot n+\phi_0)*u(n)$, where Ω_0 is the radian frequency of IF signal, T is the sampling period, and $u(n)$ is the unit step sequence. The IQ components of $f(n)$ are

$$I_0=A\cos\phi_0,\quad Q_0=A\sin\phi_0. \tag{1}$$

Fig. 2 show the IF signal and the IQ signal under the IQ coordination.[1]

The (I_0,Q_0) pair is the IQ signal we need to demodulate. Supposing (I_n,Q_n) is already known, then (I_0,Q_0) can be obtained by multiplying (I_n,Q_n) with a rotation factor $e^{-j\omega_0 n}$, where $\omega_0=\Omega_0 T$. Therefore, it is important to acquire (I_n,Q_n) at the first step. The RLC filter which is derived from the RLC resonant circuit

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[1] The sequences I_n and Q_n are the real part and the imaginary part of $Ae^{j(\phi_0+\Omega_0 T\cdot n)}$, respectively. i.e., I_n is the discrete-time IF signal and Q_n is the quadrature signal of I_n.

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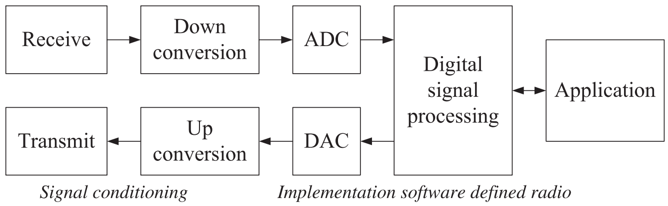

Fig. 1. Block diagram of software defined Radio system.

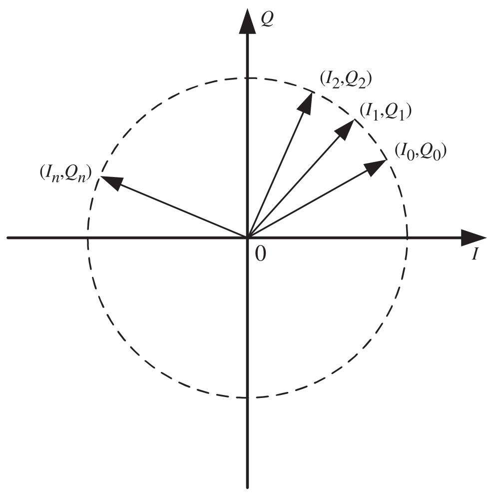

Fig. 2. IQ signal and IF signal in IQ coordinate.

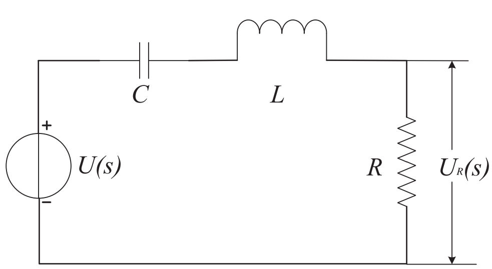

Fig. 3. Series RLC circuit.

can be used to generate (I_n,Q_n) from IF signal. It will be introduced in the next section.

3. The basic principle of RLC filter

The RLC filter is based on the RLC resonant circuit. The transfer function of a series RLC circuit (see Fig. 3) is [6]

$$H(s)=\frac{U_R(s)}{U(s)}=\frac{\frac{\Omega_0}{Q}\cdot s}{s^2+\frac{\Omega_0}{Q}\cdot s+\Omega_0^2} \tag{2}$$

where Q is the quality factor. The inverse laplace transform of Eq. (2) is the continuous-time impulse response

$$h(t)=2\alpha\left[e^{-\alpha t}\cdot\cos(\Omega_a t)-\frac{\alpha}{\Omega_a}\cdot\sin(\Omega_a t)\right] \tag{3}$$

where $\alpha=\Omega_0/2Q$ and $\Omega_a=\sqrt{\Omega_0^2-(\Omega_0/2Q)^2}$. Usually a larger value of Q is used to obtain a better frequency-selective property. Ignoring the constant multiple factor and the second small value part in the square brackets, the discrete-time impulse response can be simplified as $h(n)=e^{-\alpha T\cdot n}\cdot\Re(e^{j\omega_a n})$, where $\omega_a=\Omega_a T$. If we define

$$\mathbf{h}(n)=e^{-\alpha Tn+j\omega_a n} \tag{4}$$

then this impulse response function in complex form has two outputs. In the following we try to demonstrate that the (I_n,Q_n) pair is just involved in the steady-state response of $\mathbf{h}(n)$ for a sinusoidal input signal.

The z-transform of $\mathbf{h}(n)$ can be expressed as

$$\mathbf{H}(z)=\frac{1}{1-e^{-\alpha T+j\omega_a}\cdot z^{-1}}=\frac{z}{z-e^{-\alpha T+j\omega_a}}. \tag{5}$$

The input sinusoidal signal is $f(n)=A\cos(\omega n+\phi_0)u(n)$, the corresponding z-transform is

$$F(z)=\frac{Ae^{j\phi}}{2}\cdot\frac{z}{z-e^{j\omega}}+\frac{Ae^{-j\phi}}{2}\cdot\frac{z}{z-e^{-j\omega}}. \tag{6}$$

The output $\mathbf{Y}(z)=\mathbf{H}(z)F(z)$ is given by

$$\mathbf{Y}(z)=\frac{z}{z-e^{-\alpha T+j\omega_a}}\left(\frac{Ae^{j\phi}}{2}\cdot\frac{z}{z-e^{j\omega}}+\frac{Ae^{-j\phi}}{2}\cdot\frac{z}{z-e^{-j\omega}}\right). \tag{7}$$

There are two parts in Eq. (7), and the only difference of them is the sign of ω and ϕ. Supposing $\mathbf{M}(z)=z(z-e^{-\alpha T+j\omega_a})^{-1}\cdot z(z-e^{j\omega})^{-1}$, then $\mathbf{M}(z)$ has two poles: $z_1=e^{j\omega}$, which is on the unit circle, and $z_2=e^{-\alpha T+j\omega_a}$, which is inside the unit circle. According to the residue theorem [7,8], the inverse transform of $\mathbf{M}(z)$ is $\mathbf{m}(n)=\sum_{k=1}^{2}Res[\mathbf{M}(z)z^{n-1}]_{z=z_k}$, which is

$$\begin{aligned}\mathbf{m}(n)&=[(z-z_1)\cdot\mathbf{M}(z)z^{n-1}]_{z=z_1}+[(z-z_2)\cdot\mathbf{M}(z)z^{n-1}]_{z=z_2}\\&=\frac{e^{j\omega n}}{1-e^{-\alpha T+j(\omega_a-\omega)}}+\frac{(\exp^{-\alpha T+j\omega_a})^{n+1}}{e^{-\alpha T+j(\omega_a-\omega)}-e^{j\omega}}.\end{aligned} \tag{8}$$

When $n\to\infty$, the second part of Eq. (8) is zero, and the steady response of $\mathbf{m}(n)$ is $(1-e^{-\alpha T+j(\omega_a-\omega)})^{-1}e^{j\omega n}$. Considering the symmetry, the eventual steady response of $\mathbf{h}(n)$ is

$$\mathbf{y}(n)=\frac{Ae^{j\phi}\cdot e^{j\omega n}}{1-e^{-\alpha T+j(\omega_a-\omega)}}+\frac{Ae^{-j\phi}\cdot e^{-j\omega n}}{1-e^{-\alpha T+j(\omega_a+\omega)}}. \tag{9}$$

When $\omega=\omega_0$, considering that $\omega_a\doteq\omega_0$, Eq. (9) can be simplified as

$$\mathbf{y}(n)=\frac{Ae^{j\phi}\cdot e^{j\omega_0 n}}{1-e^{-\alpha T}}+\frac{Ae^{-j\phi}\cdot e^{-j\omega_0 n}}{1-e^{-\alpha T+2j\omega_0}}. \tag{10}$$

For k-th order $\mathbf{h}_k(n)$, which has a z-transform function $\mathbf{H}_k(z)=(1-e^{-\alpha T+j\omega_a}z^{-1})^{-k}$, the system has a multiple-order pole $e^{-\alpha T+j\omega_a}$. According to the residue theorem, the steady response of $\mathbf{h}_k(n)$ can be expressed by Refs. [7,8][2]

$$\mathbf{y}_k(n)=Ae^{j(\phi+\omega_0 n)}\left[\left(\frac{1}{1-e^{-\alpha T}}\right)^k+\left(\frac{1}{e^{-\alpha T+2j\omega_0}}\right)^k e^{-2j(\omega_0 n-\phi)}\right]. \tag{11}$$

Eq. (11) includes the IQ components and the secondary harmonic components. Normally the Q value is large enough, then the value of αT is very small, thus, the IQ components become the dominant components. Therefore, the (I_n,Q_n) pair can be obtained in the outputs of this filter.

Then the analysis of the frequency-selective properties of the RLC filter with Eq. (5) can be extended further. When $z=e^{j\omega}$, the system function $\mathbf{H}(z)$ becomes the frequency response $\mathbf{H}(e^{j\omega})$. Defining the relative power spectral density function $P_1(e^{j\omega})=\left|\mathbf{H}(e^{j\omega})/\mathbf{H}(e^{j\omega_0})\right|^2$. Considering that $\omega_a\doteq\omega_0$, then $P_1(e^{j\omega})$ can be simplified as[3]

$$P_1(e^{j\omega})=\frac{1}{1+\left(\frac{\omega-\omega_0}{\alpha T}\right)^2}. \tag{12}$$

[2] Some comparisons of several different kinds of high order infinities are involved in the derivation of Eq. (11).

[3] Taylor series expansion is applied in this derivation.