A NOVEL IF CHANNEL SELECTION TECHNIQUE BY ANALOG-DOUBLE QUADRATURE SAMPLING FOR COMPLEX LOW-IF RECEIVERS

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***Abstract*—This paper presents a novel IF channel selection technique for complex low-IF receivers. By utilizing the frequency-shifting characteristic of analog-double quadrature sampling (A-DQS), and setting the IF equal to half of the channel bandwidth, the desired channel and its image are neighbors and they can be selected to the baseband through simple control. Thus, the channel selection original performed by the front-end integer-N PLL frequency synthesizer can be partitioned to the A-DQS in the IF. The benefits of this partitioning channel selection method are 1) the step-size of the integer-N PLL frequency synthesizer can be doubled to two channels bandwidth in order to maintain low phase noise and fast settling time in very low-IF operation, 2) the locking position of the local oscillator in the entire frequency band is halved, since the frequency synthesizer only performs the frequency down-conversion of every two channels to the IF. Such technique is especially efficient in wireless receivers for 2.4GHz 802.11b WLANs or Bluetooth that employ frequency-hopping to capture the signals.**

I. INTRODUCTION

Today's high demand of low-power and high-integrated wireless receivers implies that IF receiver topologies like heterodyne [1] or super-heterodyne [2] are no longer adequate as they require several off-chip components. Therefore, only high integratability receiver topologies like zero-IF [3] and complex low-IF [4-5] emerge as the solutions. For zero-IF topology, even if it possesses simple structure and low-power consumption, the dominant problems like DC-offset, even-order distortion and *1/f* noise limit its performance. However, such problems are insensitive in complex low-IF topology, since the desired signal is offset from the low frequency noise corner. Moreover, since low-IF receiver topology tackles the image problem through signal cancellation on chip, it also possesses high degree of integration as the zero-IF one. Therefore, in this paper, a novel channel selection technique will be presented for a complex low-IF receiver as shown in Fig.1, based on the analog-double quadrature sampling (A-DQS) [6-7]. When the intermediate frequency (IF) is set as the lowest possible value (half of the channel bandwidth) to minimize the power consumption and the image rejection requirement, the extra benefits are that the A-DQS can perform IF-to-baseband frequency down-conversion, as well as, channel selection between two adjacent channels through simple control. Thus, the function of channel selection can be partitioned from the front-end integer-N PLL frequency synthesizer to the A-DQS at the IF, to enhance the entire receiver performance.

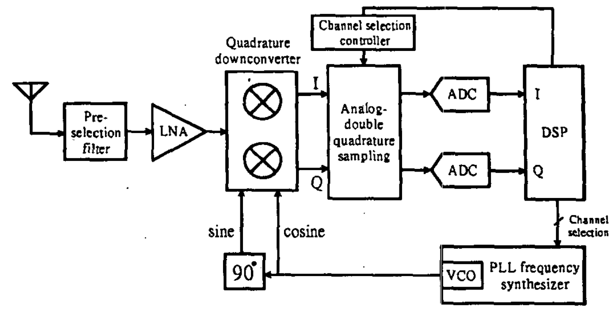

Fig. 1 Proposed complex low-IF receiver topology

This paper will first describe the proposed channel selection technique in section II. Simulation results are provided in section III. A comparison between the traditional and proposed frequency down-conversion approaches is addressed in section IV. Section V presents the conclusion.

II. CHANNEL SELECTION BY A-DQS

In order to fully illustrate how the frequency shifting property of A-DQS can perform the function of channel selection between two adjacent channels, the mathematical model of A-DQS will be presented in this section. The front-to-back end frequency down-conversion and channel selection are described in Fig. 2. The band-limited RF channels (labeled **A**, **B**, **C** and **D**) in Fig. 2a are first filtered and then amplified by a pre-selection filter and a LNA, respectively. It can be considered also that the desired channels are **A** and **B**, given by:

$$x_{RF}(t) = A(t)\cos[\omega_{SIG}t + \phi_A(t)] + B(t)\cos[\omega_{IMG}t + \phi_B(t)] \quad (1)$$

For coherent detection receivers, equation (1) is conveniently re-expressed with two components, in-phase (*I*) and quadrature-phase (*Q*) given by:

$$x_{RF}(t) = I_A(t)\cos(\omega_{SIG}t) - Q_A(t)\sin(\omega_{SIG}t) + I_B(t)\cos(\omega_{IMG}t) - Q_B(t)\sin(\omega_{IMG}t) \quad (2)$$

where

$$I_A(t) = A(t)\cos\phi_A(t), Q_A(t) = A(t)\sin\phi_A(t)$$
$$I_B(t) = B(t)\cos\phi_B(t), Q_B(t) = B(t)\sin\phi_B(t) \quad (3a\text{-}d)$$

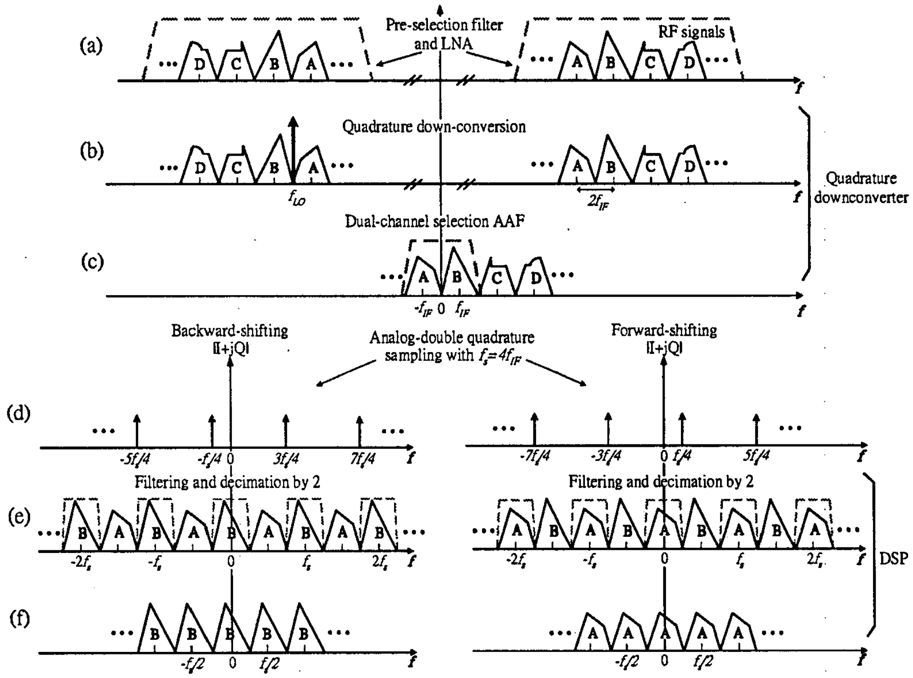

Fig. 2 Proposed front-to-back end frequency down-conversion method and channel selection by A-DQS

The dual channel **A** and **B** are then mixed down to the IF by a quadrature downconverter driven by the local oscillator (LO). The frequency location of the LO is selected between every two channels, such as channel **A** and **B** (Fig. 2b). Thus, channel **B** will be the corresponding image of **A** or vice versa. The key benefit of this IF operation is the low image rejection requirement as the maximum power of the adjacent channel is much smaller than the other in-band channels in most wireless communications. The down-converted channels **A** and **B** after lowpass filtering by an anti-aliasing filter (AAF), are settled at $-f_{IF}$ and $+f_{IF}$, respectively, as given by:

$$\begin{aligned} x(t) &= x_I(t) + jx_Q(t) \\ &= \frac{I_A(t)}{2}e^{-j(\omega_{LO}t+\phi_{LO})} + j\frac{Q_A(t)}{2}e^{-j(\omega_{LO}t+\phi_{LO})} \\ &+ \frac{I_B(t)}{2}e^{j(\omega_{LO}t+\phi_{LO})} + j\frac{Q_B(t)}{2}e^{j(\omega_{LO}t+\phi_{LO})} \end{aligned} \tag{4}$$

Next, the signals will face a complex sampling operation (starting the A-DQS) that will result in a forward and backward shifting, as shown in Fig. 2d, with orthogonal samplers $P_I(t)$ and $P_Q(t)$,

$$P_I(t) = \sum_{n=-\infty}^{\infty} [\delta(t-nT_s) - \delta(t-nT_s-T/2)]$$
$$P_Q(t) = \sum_{n=-\infty}^{\infty} [\pm\delta(t-nT_s-T/4) \mp \delta(t-nT_s+T/4)] \tag{5a-b}$$

with sampling frequency, $f_s=4f_{IF}$ as shown in Fig. 3.

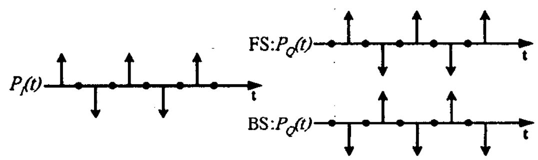

Fig. 3 Complex samplers in time domain: $P_I(t)$, FS:$P_Q(t)$ (forward shifting) and BS:$P_Q(t)$ (backward shifting)

For forward shifting (FS), it will be necessary to consider the equation (5b) with the upper sign, whereas for backward shifting (BS), the lower sign must be considered in the same equation, and the same rule applies in the following equations.

On the other hand, in the frequency domain, the Fourier transforms of $P_I(t)$ and $P_Q(t)$ are

$$P_I(j\omega) = \sum_{k=-\infty}^{\infty} 2\pi a_k \delta(\omega - k\omega_{IF})$$
$$P_Q(j\omega) = \sum_{k=-\infty}^{\infty} 2\pi b_k \delta(\omega - k\omega_{IF}) \tag{6a-b}$$

which they are shown in Fig. 4 for FS and BS conditions, where a_k and b_k are the Fourier coefficients,

$$a_k = \begin{cases} \frac{2}{T} & for\ k = 2n+1, n = 0,\pm1,\pm2,\cdots \\ 0 & otherwise \end{cases}$$
$$b_k = \begin{cases} \frac{2}{T}e^{\mp jk\pi/2} & for\ k = 2n+1, n = 0,\pm1,\pm2,\cdots \\ 0 & otherwise \end{cases} \tag{7a-b}$$

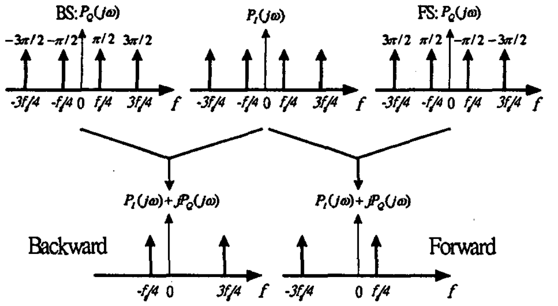

Fig. 4 Frequency domain of samplers $P_I(j\omega)$, FS:$P_Q(j\omega)$, BS: $P_Q(j\omega)$ and $P_I(j\omega)+j\,P_Q(j\omega)$ for FS and BS

The complex sampler $P\,(j\omega)$ is obtained by combining $P_I\,(j\omega)$ and $P_Q\,(j\omega)$, as shown in Fig. 4 also, and is given by:

$$P(j\omega) = P_I(j\omega) + jP_Q(j\omega)$$
$$= \sum_{k=-\infty}^{\infty} 2\pi c_k \delta(\omega - k\omega_{IF}) \tag{8}$$

where c_k is the Fourier coefficient,

$$c_k = a_k + jb_k$$
$$= \begin{cases} 4/T & for\ k = 4n \pm 1, n = 0, \pm 1, \pm 2, ... \\ 0 & otherwise. \end{cases} \tag{9}$$

Thus, the complex sampling operation is equivalent to a multiplication of the output by the complex term,

$$\cos(n\pi/2) \pm j\sin(n\pi/2) \quad n = 0,1,2,... \tag{10}$$

leading to the sampled output (Fig. 2f) represented by:

$$y(nT) = [x_I(nT)\cos(n\pi/2) \mp x_Q(nT)\sin(n\pi/2)] \quad n = 0,1,2,...$$
$$+ j[x_Q(nT)\cos(n\pi/2) \pm x_I(nT)\sin(n\pi/2)] \tag{11}$$

Equation (11) can be easily demonstrated in the time domain as shown in Fig. 5. Suppose, in forward shifting, two single-harmonic input signals

$$x(t) = A(t)e^{-j[\omega_{IF}t+\phi_A(t)]} + B(t)e^{j[\omega_{IF}t+\phi_B(t)]} \tag{12a}$$

$$x_I(t) = \mathrm{Re}\{x(t)\} = A(t)\cos[\omega_{IF}t + \phi_A(t)]$$
$$+ B(t)\cos[\omega_{IF}t + \phi_B(t)] \tag{12b}$$

$$x_Q(t) = \mathrm{Im}\{x(t)\} = -A(t)\sin[\omega_{IF}t + \phi_A(t)]$$
$$+ B(t)\sin[\omega_{IF}t + \phi_B(t)] \tag{12c}$$

that are sampled-and-held by the A-DQS. Such IF inputs $+IF_i$, $-IF_q$, $-IF_i$, $+IF_q$ are sampled by I channel during clock phases, 1,2,3 and 4 respectively; $+IF_q$, $+IF_i$, $-IF_q$, $-IF_i$ are sampled by Q channel during clock phases, 1,2,3 and 4 respectively. On the other hand, in the frequency domain, the channel **A** is obtained at $\pm nf_s$ for $n=0,1,2...$ whereas channel **B** (image of **A**) is shifted to $\pm nf_s/2$ for $n=1,3,5...$

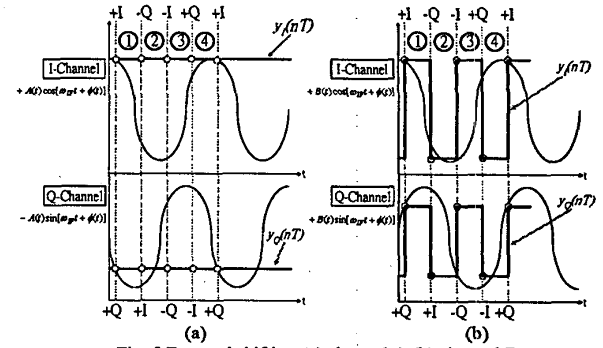

Fig. 5 Forward shifting (a) channel A (b) channel B

Similar results can be obtained for backward shifting. Clearly, both cases do not suffer from the DC offset and $1/f$ noise problems since those noises are shifted to $\pm nf_s/4$, for $n=1,3,5....$ Finally, after complex sampling through A-DQS followed by A/D conversion, the I and Q data at a rate $f_s/2$ can be obtained through digital decimation by a factor of two, as shown in Fig. 2e and f.

As illustrated above, the channel selection can be effectively realized by exchanging the sampling sequence between clock phases 2 and 4, as shown in Fig. 6. Thus, a simple digital circuit (channel selection controller) controlled by the DSP can be employed to perform such function.

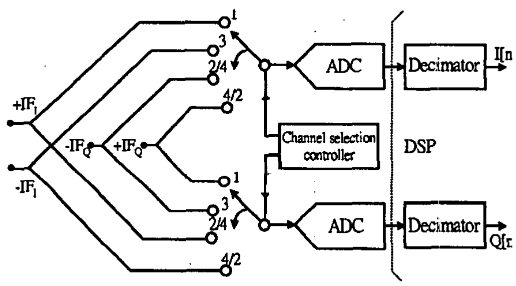

Fig. 6 A-DQS channel selection topology

III. SIMULATION RESULTS

The A-DQS scheme and two 8 bits Nyquist rate A/D converters were modeled at the system level in a *Matlab™* environment and simulated with *Simulink™*. Three test tones: channel **A** (image of channel **B**), channel **B** (image of channel **A**) and a zero frequency tone are applied in the simulation as shown in Fig. 7a. The zero-frequency component is applied to test the low frequency noise sensitivity. To acquire the channel **A** in baseband, forward shifting is used and the power spectrum is shown in Fig. 7b. Channel A is shifted to $\pm nf_s$ for $n=0,1,2,$ ... while the zero-frequency tone and channel B are shifted to $\pm nf_s/4$ and $\pm nf_s/2$ for $n=1,3,5...$, respectively. A similar result is shown in Fig. 7c for backward shifting. Therefore, all the results are consistent with the theoretical analysis presented in the previous section.

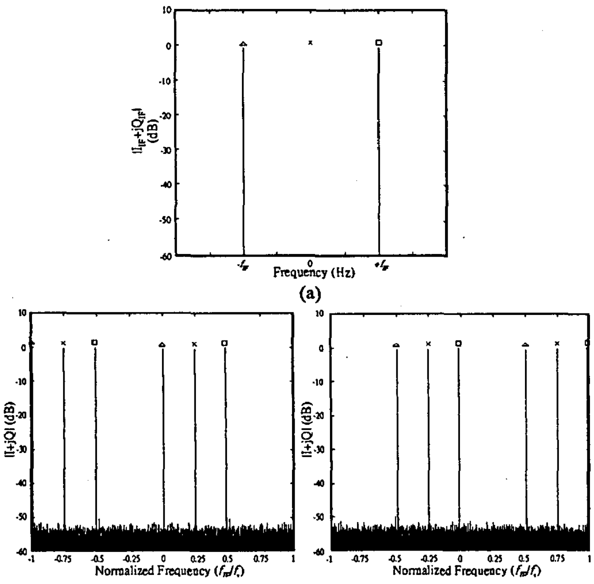

Fig. 7 Simulated power spectrum of (a) inputs (b) forward-shifted outputs (c) backward-shifted outputs (Δ: channel **A**, □: channel **B**, ×: zero-frequency component)

IV. Comparison between Traditonal and Proposed Frequency Down-Conversion Methods

In nominal low-IF receivers, the desired channel is mixed down in the way shown in Fig. 8a. The required step-size is conveniently set to one channel bandwidth (BW), to ease the channel selection by utilizing simple integer-N PLL frequency synthesizer. The required locking positions of the local oscillator (LO) are equal to the number of channels in the whole frequency band.

Differently, in the proposed method, since the A-DQS can perform channel selection between two channels in the IF, the locking position of the LO can be applied between every two channels as shown in Fig. 8b. For instance, channels **A** and **B** are down converted together by the (LO) frequency f_{n-1}. Similarly, channels **C** and **D** can be downconverted by the next frequency f_n. With this method, first, there are no DC-offset and $1/f$ noise problems since their position in the frequency spectrum is shifted to $fs/4$ for $n=0,1,2,...$ prior to the A/D conversion. Second, the minimum frequency resolution of the frequency synthesizer can then be extended from one to two channels BW, since the locking position is selected on every two BWs. Third, the total required number of locking position is halved as the final selection between the dual-channel is performed by the A-DQS. For wireless communications such as Bluetooth, there would be a total of 79 channels with BW=1MHz per channel. The frequency synthesizer needs to perform frequency hopping in at least 75/79 channels pseudo-randomly with 625μs/hop residence time.

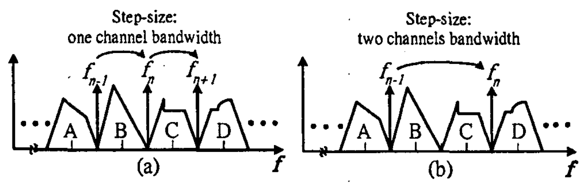

Fig. 8 Frequency down-conversion (a) nominal method and (b) proposed method

By using the proposed technique with an integer-N PLL frequency synthesizer, the locked position of the LO requires only 40 channels, and the step-size can be increased from 1MHz to 2MHz. Thus, reducing the problems of large phase noise and long settling time of the frequency synthesizer in very small step-size operation, due to insufficient bandwidth for the PLL.

V. Conclusion

We have presented a novel IF channel selection technique for complex low-IF receivers by utilizing an analog-double quadrature sampling scheme. By employing such technique, even if the frequency synthesizer operates in very low intermediate frequency, the step-size of the frequency synthesizer is kept high to maintain low phase noise and fast settling time. Moreover, the required number of locking positions of the local oscillator is halved as the duty of channel selection is now shared with the A-DQS from the RF to the IF. Such scheme is verified by computer simulation and its application is focus on wireless receivers, which need to perform frequency-hopping repeatedly.

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