DWA Algorithms for Multibit Complex Bandpass ΔΣAD Modulators of Arbitrary Signal Band

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ABSTRACT

This paper describes Data-Weighted-Averaging (DWA) algorithms for multibit complex bandpass ΔΣAD modulators with arbitrary signal bands. In most of bandpass ΔΣAD modulators, their sampling frequency is chosen as four times of the center frequency of their signal band in order to simplify the following digital filter design. But with this relationship, their performance would be limited by the third harmonics of the input signal because it is aliased into the signal band; hence recently some bandpass ΔΣAD modulators whose the sampling frequency is not at four times of their signal center frequency are proposed to avoid this problem. Our proposed DWA algorithms are applicable to such multibit complex bandpass ΔΣAD modulators whose sampling frequency is not necessarily at four times of the signal center frequency; these algorithms are realized by applying our previously developed multi-bandpass (N-path) DWA algorithms to 2-channel DACs in complex modulators. We have verified their operation by Matlab simulation.

Keywords: *Comlplex ΔΣ Modulator, Multi-bit, DWA Algorithm, Error Correction, Noise Shaping*

1. Introduction

In the RF receiver of communication systems such as cellular phones and wireless LANs, low-IF receiver architecture is frequently used so that more receiver functions, such as multi-standard and automatic gain control, can be moved to the digital part to provide more programmability. In conventional low-IF receiver architectures, two real (one input and one output) ΔΣAD modulators are used for In-phase (I) and Quadrature (Q) paths. Its disadvantage is that not only input signals but also image signals are converted by ADCs. On the other hand, a complex bandpass ΔΣAD modulator can provide superior performance to a pair of real bandpass ΔΣAD modulators of the same order. It performs AD conversion only for the positive frequency of I, Q input signals not for the negative frequency (image signal), and AD conversion can be realized with low power dissipation. So they are desirable for such low-IF receiver applications[1]$^{-}$[4]. Most art of the design of bandpass ΔΣAD modulators can be devrived from sampling frequency of modulator at $f_s = 4f_{in}$ (f_{in}: the center frequency of input signal), in order to make the following decimation filter simple[6]. However, the mismatch between the I, Q paths and harmonic distortion of input signal in the complex bandpass modulator cause image signal, and any tone due to mirrior image is in the band of intreset with the closest one at $f_s/4$, which degrades the SNDR of modulator. The mirror image free technique is proposed to avoid the influence of this harmonic distortion[5]. It chooses f_s at different frequency from $4f_{in}$, so that the image tones closest at $f_s/4$ are located outside of the band of interest. (This technique made the following digital filter becomes complicated, but the cost and performance of digital circuits is not a dominant problem in CMOS process technology.) Moreover, the use of a low-order multibit ΔΣAD modulator makes higher SNDR possible with lower order of loop-filters, and the stability problem is alleviated. It is attractive for low power implementation because it alleviates the slew-rate requirements of operational amplifiers with high dynamic range in the modulator. However, multibit DACs cannot be made perfectly linear and their nonlinearity in the feedback paths are equivalent to errors added directly to the input signals, and hence they may degrade the SNDR of the ΔΣAD modulator[6]. We have developed a DWA algorithm for complex bandpass modulators which makes the nonlinearities of DACs noise-shaped at $f_{in} = f_s/4$ [7, 8]. However, proposed algotrithm only suppress nonlinearity effects of multibit DACs at the case of $f_s/4$, and is not avilable for other cases. This paper presents new DWA algorithms for multibit complex bandpass ΔΣ AD modulators whose the sampling frequency is not necessarily at four times of their signal center frequency, these algorithms are

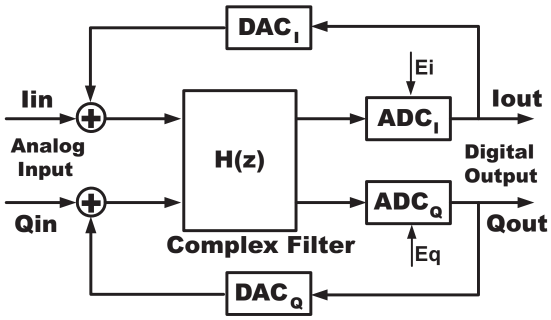

Fig. 1. Complex bandpass ΔΣAD modulator block diagram.

realized by applying our previously developed multi-bandpass (N-path) DWA algorithms[9] to 2-channel DACs in complex modulators. We have conducted MATLAB simulations to confirm the effectiveness of the proposed algorithms.

2. Multibit Complex Bandpass ΔΣ AD Modulator of Arbitrary Signal Band

2.1 Complex Bandpass ΔΣAD Modulator

Fig.1 shows the simplified block diagram of the complex bandpass ΔΣAD modulator; which is composed of a complex bandpass filter, two internal quantizers(ADCs) and two DACs. When input signal $X(z)$, output signal $Y(z)$ and quantizer noise $E_q(z)$ are given by complex form,

$$\begin{aligned} X(z) &= I_{in} + jQ_{in} \\ Y(z) &= I_{out} + jQ_{out} \\ E_q(z) &= E_i + jE_q \end{aligned}$$

we can get the relationship of input and output of this complex modulator as following:

$$I_{out} + jQ_{out} = \frac{H(z)}{1+H(z)}(I_{in} + jQ_{in}) + \frac{1}{1+H(z)}(E_i + jE_q) \quad (1)$$

Here, $H(z)$ is a complex filter transfer function, then we have signal transfunction $STF(z)$ and noise transfuntion $NTF(z)$ as followings:

$$STF(z) := \frac{H(z)}{1+H(z)} \quad (2)$$

$$NTF(z) := \frac{1}{1+H(z)} \quad (3)$$

We see that complex bandpass ΔΣAD modulator with two inputs and two outputs of I and Q, two analog input signals are modulated in complex form, and gets two digital output signals. Quantization noise of two ADCs $E(z) = E_i + jE_q$ is noise shaped in complex form by NTF(Eq.3) of ΔΣ modulator. Compex bandpass filter in the modulator has asymmetrical frequency characteristics to the axis of $\omega = 0$, which is different from a real bandpass filter. It has opposite frequency characteristics for $\omega > 0$ and $\omega < 0$, one side is signal-band (passband), another side is image-band (attenuation band). Therefore, a complex bandpass ΔΣAD modulator performs AD conversion effectively only for the positive frequency of I, Q input signals in a low-IF receiver, and hence it can be realized with lower power dissipation than a pair of real bandpass ΔΣAD modulators which perform AD conversion for the negative frequency (image signal) as well as the positive frequency .

2.2 Bandpass ΔΣAD Modulator of Arbitrary Signal Band

The simplest way to design of a bandpass ΔΣAD modulator is to choose $f_s = 4f_{in}$ in order to simplify the following digital filter design[6]. With this relationship, the odd harmonic distortions of the input signal is aliased into the signal band of intreset as an image signal, whose frequency is closest one at $f_s/4$; and hence the performance of modulator would be damaged. As shown in Fig.2(a), we assume that the frequency of input signal is $f_{in} = f_s/4 - \Delta f$, then the frerquency of third harmonic of input signal is $HD_{3^{rd}}(f_{in}) = 3f_s/4 - 3\Delta f$; this harmonic is mirrored by $f_s/2$, which appears at the frequency of $f_s/4 - 3\Delta f$. This image signal is close to the input signal in the band of intreset, which degrades the SNDR of modulator. Actually, not only third harmonic distortion but also other odd harmonic distortions such as fifth and seventh harmonic distortions exist in modulator circuits too, they will be aliased into the signal band as the same mechanism.

A mirror image free technique is proposed and implemented to avoid the influence of harmonic distortion for real (one input and one output) bandpass ΔΣAD modulators[5]. This method is effective with complex (two inputs and two outputs) ΔΣAD modulator too. As shown in Fig.2(b), we chose f_s at different frequency from $4f_{in}$, for example while $f_{in} = f_s/6$, then the image tone of third harmonic appears at $3\Delta f$ from $f_s/4$, which is near to $f_s/2$ in this case. Obviously that there is not influence to

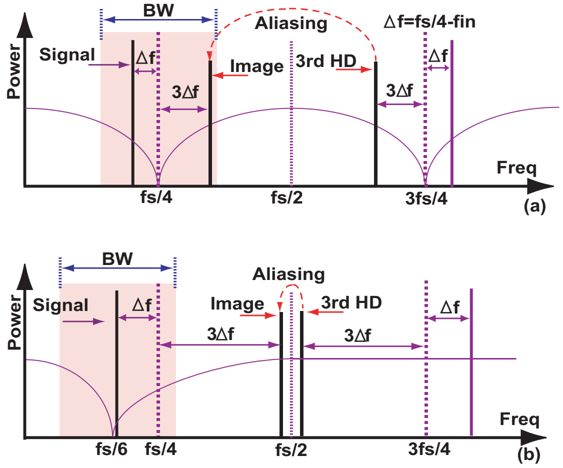

Fig. 2. (a) Third harmonic distortion aliasing into the signal band. (b) Third harmonic distortion aliasing out of the signal band.

SNDR any more because the aliased tones are located at outside of the band of interest.

2.3 Nonlinearity of DAC in multi-bit ΔΣAD modulator

Multi-bit ADCs/DACs are used inside the ΔΣAD modulator to obtain higher performance with lower power dissipation. The quantization error is reduced by 6dB for every bit added to the resolution of the quantizer. Smaller steps of quantizer result in a lower quantization error, feedback loop becomes more linear, and the stability problem is alleviated. Larger input signals can be allowed, higher SNDR is possible for low order of loop filter with lower OSR. Input signal of loop filter changes less, the required slew rate of the input operational amplifiers of loop filter is relaxed, it is attractive for low power implementation. However, multibit DACs cannot be made perfectly linear and their nonlinearity in the feedback paths are equivalent to errors added to the input signals, will appear directly at output, and hence they may degrade the SNDR of the ΔΣAD modulator.

Some signal processing algorithms such as DEM (Dynamic Element Matching) and DWA techniques are implemented to suppress nonlinearity effects of DACs, which improve the SNDR of multi-bit ΔΣAD modulator. However, either of them is for real ΔΣAD modulators (single-input & output only)[10]. We have proposed a complex bandpass DWA algorithm which noise-shapes nonlinearities of multi-bit DACs in a complex bandpass AD modulator which

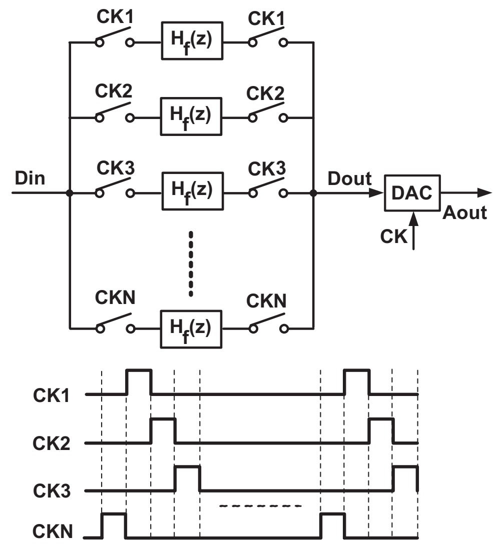

Fig. 3. Proposed architecture of arbitrary signal band DWA algorithms.

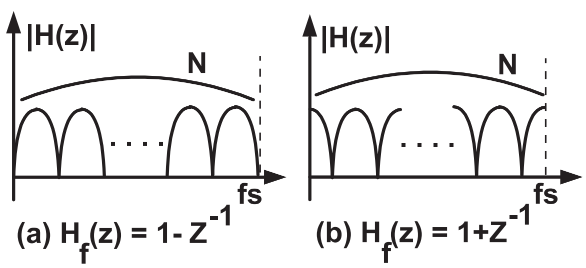

Fig. 4. Frequency response of N-path filter.

has I, Q inputs and outputs, but it is suitable for complex bandpass ΔΣAD modulator in case of its sampling frequncy is $f_s = 4f_{in}$ only. This paper presents algorithms for complex bandpass modulator of arbitrary signal band (arbitrary sampling frequency), and they noise-shape nonlinearities of multi-bit DACs in complex modulator.

3. Arbitrary Signal Band DWA Algorithms

We have proposed a multi-bandpass DWA algorithm with one-input and one-output[9], then we provide it to two DACs for noise-shaping their nonlineaity in complex bandpass ΔΣAD modulator. Fig.3 shows proposed architecture of arbitrary signal band DWA algorithms with one-input and one output. We consider a digital filter is added to the fornt-end of DAC here, and the sampling frequency of DAC, CK is f_s, and the frequency of $CK1, CK2, CK3, ...CKN$ are same as f_s/N. The filter at front-end of DAC is a interleaved architechture of $H_f(z)$, and We apply two different types of transfer function to $H_f(z)$ as following:

3.1 DWA Algorithm I: $H_f(z) = 1 - z^{-1}$

When the transfer function of digital fiter at fornt-end of DAC is written by $H_1(z) = 1 - z^{-N}$, the frequency response of magnitude characteristics for $H_1(z)$ is shown as Fig.4(a). The zeros of $H_1(z) = 0$ (plural center of IF signals in the modulator) are f_n, which given from $z = \exp(j2\pi f)$ can be written as following:

$$f_n = \frac{n}{N} f_s.$$

Here, $n = 0, 1, 2, 3, ...$ $(n < N)$. Then we can get $N+1$ notches for noise-shaping at signal band while we arrange $N+1$ zeros of $H_1(z)$ at frequency axis from DC to f_s. DC is one of zeros in this case, above metioned algorithm becomes a LP DWA algorithm to noise-shape DC signal while $N = 1$.

3.2 DWA Algorithm II: $H_f(z) = 1 + z^{-1}$

When the transfer function of digital fiter at fornt-end of DAC is written by $H_2(z) = 1 + z^{-N}$, the frequency response of magnitude characteristic for $H_2(z)$ is shown as Fig.4(b). Plural center of IF signals in the modulator, f_n are give by $H_2(z) = 0$, then we get following:

$$f_n = \frac{2n+1}{2N} f_s.$$

Here, $n = 0, 1, 2, 3, ...$ $(n < N)$. Then we can get N notches for noise-shaping at signal band while we arrange N zeros of $H_2(z)$ from DC to f_s.

We can get noise-shaping of nonlineaty in multi-bit DAC for arbitrary signal band. We also can reduce influence of a nonlinearity of DACs while we apply above algorithms to two multi-bit DACs in complex modulator of arbitrary signal band.

4. Realization of Arbitrary Signal Band DWA Algorithms

The direct realization of the above filters architecture in Fig.4 are *NOT* possible, then we proposed the equivalent realization of the digital filters at the fornt-end of DAC.

We modify the two segmented switched-capacitor (SC) DACs as following:

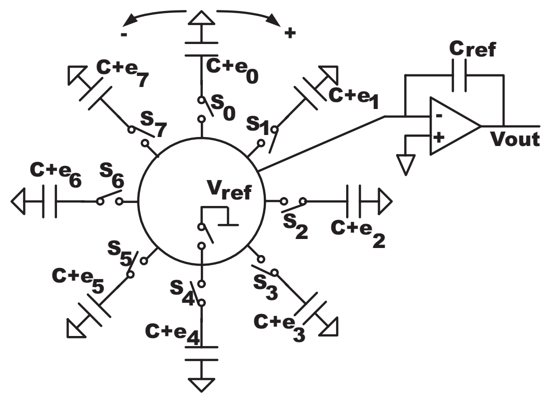

Fig. 5. Switched-capacitor segmented DAC in a ring form.

- Arrange unit-capacitor-cell of each DAC in a ring form as shown in Fig.5.
- Consider the direction for arranged capacitor-cell. For each DAC in a ring form, the unit-capacitor-cells should be selected to turn ON according to the iuput signal with bi-directional rotation of + and− in decided turn.
- Let each DAC have a pointer to show which unit-current-cell should be selected next sample time, and let the pointers for DAC1 and DAC2 be $P_1(n)$ and $P_2(n)$ at time n.

4.1 Realization of $H_1(z) = 1 - z^{-N}$

Consider the front-end filter of DAC shonw in Fig.3 in case of the transfer function is given by $H_f(z) = 1 - z^{-1}$. We know that $H_f(z) = 1 - z^{-1}$ can be realized in lowpass DWA rotation manner, and hence $H_1(z)$ can be considered as ⌈N simple lowpass DWA algorithms with N pointers are interleaved by N times ⌋. Therefore we get the realization of noise-shping for nonlinearity of DAC by $1 - z^{-N}$. Fig.6 shows an operation example while $N = 4$, the unit-capacitor-cells in ON state which are filled in black when input data are sequentially given by 4, 3, 2, 2, 5, 3, 4, 6.

4.2 Realization of $H_1(z) = 1 + z^{-N}$

Consider the front-end filter of DAC shonw in Fig.3 in case of the transfer function is given by $H_f(z) = 1 + z^{-1}$. We know that $H_f(z) = 1 + z^{-1}$ can be realized in highpass DWA rotation manner, and hence $H_2(z)$ can be considered as ⌈N simple highpass DWA algorithms with N pointers are interleaved by N

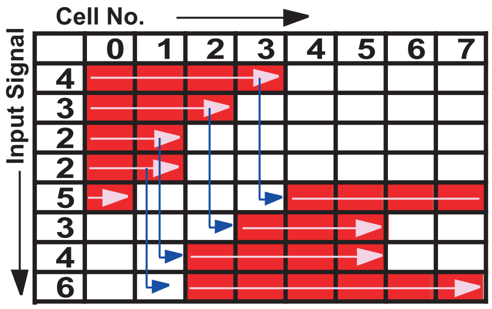

Fig. 6. Arbitrary signal band DWA algorithm I operation example for N = 4, here $H_1(z) = 1 - z^{-4}$.

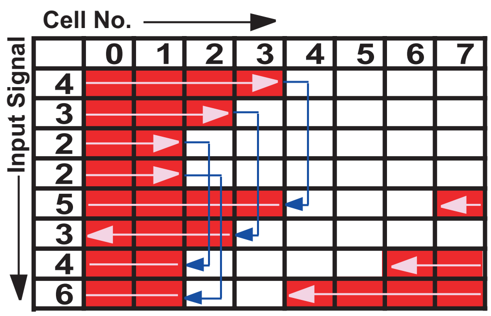

Fig. 7. Arbitrary signal band DWA algorithm II operation example for N = 4, here $H_2(z) = 1 + z^{-4}$.

times ⌋. Therefore we get the realization of noise-shping for nonlinearity of DAC by $1 + z^{-N}$. Fig.7 shows a operation example while $N = 4$, the unit-capacitor-cells in ON state which are filled in black when input data are sequentially given by 4, 3, 2, 2, 5, 3, 4, 6.

5. Simulation Results

We have performed MATLAB simulations to confirm the noise-shaping function on multi-bit DAC non-linearities of the proposed algorithms in a arbitrary signal band complex ΔΣAD modulator.

Fig.8 shows block diagram of simulated complex bandpass ΔΣAD modulator. The STF and NTF of complex modulator are given by following:

$$\begin{aligned} STF(z) &= z^{-2} \\ NTF(z) &= z^{-2}\Big(z - \big(\frac{-\sqrt{2}}{2} + j\frac{\sqrt{2}}{2}\big)\Big)^2 \end{aligned}$$

Fig.9 shows output power spectrum of the modulator, and we see that signal center frequency is $3f_s/8$.

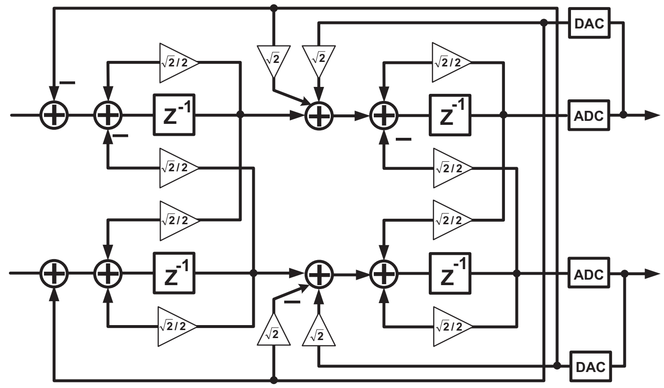

Fig. 8. Block diagram of complex bandpass ΔΣAD modulator.

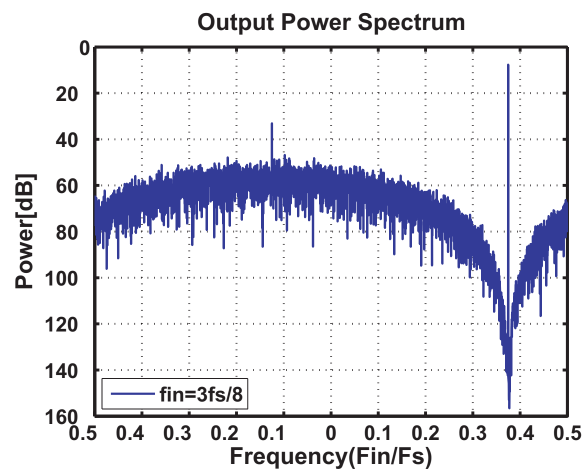

Fig. 9. Simulation result for output power spectrum of complex bandpss ΔΣAD modulator.

In our simulation, the same seconde-order complex bandpass filter and two ideal internal 9-level ADCs in a ΔΣ modulator were used. However two 9-level DACs in a modulator in four cases are different as follows:

Case 1 Two identical and ideal (linear) DACs were used.

Case 2 Two normal segmented SC DACs with mismatches among unit-capacitor-cells were used.

Case 3 Two DACs whose mismatches are the same as case 2 and which employ the proposed DWA algorithm I were used.

Case 4 Two DACs whose mismatches are the same

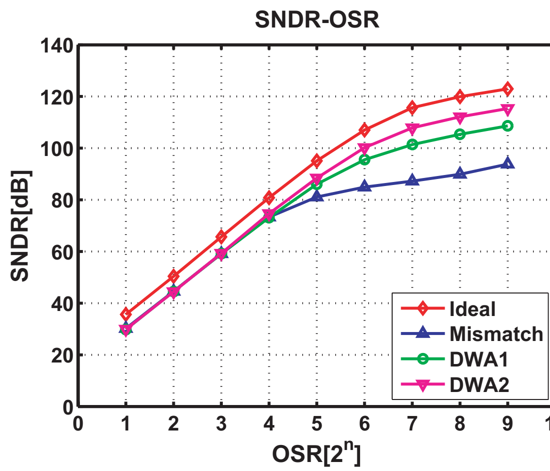

Fig. 10. Simulated results comparison for SNDR vs OSR in compex bandpass ΔΣAD modulator.

as case 2 and which employ the proposed DWA algorithm II were used.

Fig.10 shows the simulation results comparison for SNDR versus OSR of modulators in four cases. We see that in case 1, SNDR of the modulator increases as OSR increases. However in case 2, SNDR saturates even OSR increases; On the other hand, in cases 3 and 4, the noise caused by nonlinearities of DAC is pushed out from the signal band, and their influence to the modulator accuracy is reduced which leads to the SNDR improvement. Moreover, to get the noise-shaping at the signal center frequency of $3f_s/8$, 8-order filter is necessary while DWA algorithm I is used, however, only 4-order filter is necessary while DWA algorithm II is chosen. Therefore, we got that in case of signal center frequency is $f_n = (2n+1)f_s/2N$, DWA algorithm II is preferable to DWA algorithm I for half the order of filter and circuits, and hence it can be realized simpler.

6. Conclusion

We have proposed two DWA algorithms for noise-shaping multi-bit DAC nonlinearities in complex bandpass ΔΣAD modulator of arbitrary signal band. These algorithms are realized by applying our previously developed multi-bandpass (N-path) DWA algorithms to 2-channel DACs in complex modulators.

Our proposed DWA algorithms are applicable to such multibit complex bandpass ΔΣAD modulators whose sampling frequency is not necessarily at four times of the signal center frequency, so that the influence from harmonic distortions is avoided, and SNDR deterioration is improved. The effectiveness of proposed algorithms are confirmed by MATLAB simulation.

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