A Mixed-Signal MIMO Beamforming Receiver

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Abstract—This paper presents a mixed signal MIMO beamforming architecture based on the oversampling and coarse quantization. Due to the strenuous dynamic range and power consumption requirements on analog-to-digital converters (ADC) in fully digital implementations of MIMO systems, a mixed-mode implementation is examined. To achieve an analog implementation of a MIMO beamforming receiver, simultaneous gain and phase-shift control in each receive path is needed. The architecture introduced in this paper both relaxes analog circuit requirements and accomplishes simultaneous gain and phase control for each antenna.

Index Terms—Vector modulators, RF VGA, MIMO systems, CMOS integrated circuits, mixed analog-digital integrated circuits

I. INTRODUCTION

Multiple-input multiple-output (MIMO) antenna systems are a central component in many next generation air interface standards due to the multiplexing and diversity gains that they enable. Current digital MIMO system implementations typically contain separate down-conversion paths and analog-to-digital converters (ADC), and signal combining is done in the digital domain after conversion [1]. Retaining a separate ADC per down-conversion path not only increases the dynamic range requirement of each ADC, but also the power consumption of the entire system. This can be avoided if the signals are combined with circuitry in the analog domain.

To perform beamforming, the receiver must precisely scale the amplitude of each received signal in addition to shifting its phase. This can be performed in the analog domain by a vector demodulator. Vector demodulators can be realized with RF variable gain amplifiers (VGA), but fine-resolution VGAs are typically implemented in non-CMOS processes such as GaAs [2].

Rather than implementing these variable gains and phase-shifts with complex analog circuitry, we introduce a mixed-mode vector demodulator architecture based on oversampling and coarse quantization. While $\Sigma\Delta$ ADCs achieve high resolution by supplementing coarse quantization with oversampling in the time domain, we borrow this concept and achieve increased phase-shift and gain resolution by coarse quantization and oversampling in the phase domain. The proposed architecture translates most of the system complexity from the analog domain to the digital domain and realizes the vector demodulation functionality with simple analog circuits.

II. MIMO BEAMFORMING

The signals impinging on a set of receive antennas can be written as

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{z} \quad (1)$$

where x_m , the m th component of $\mathbf{x}$, is the transmitted signal from the m th transmit antenna, y_n , the n th component of $\mathbf{y}$, is the received signal from the n th receive antenna, h_{mn} is the gain between the m th transmit antenna and the n th receive antenna, and z_n is the additive noise plus interference from the n th receive antenna [3]. The components of the channel matrix $\mathbf{H}$ are usually modeled as random variables.

To recover the original transmitted signal, a complex multiplication must be performed on each component of $\mathbf{y}$, and the products must be added coherently at the output. This may be written as

$$r = \mathbf{v}^\dagger \mathbf{y} \quad (2)$$

$$\mathbf{v} = \begin{bmatrix} a_0 e^{-j\phi_0} \\ a_1 e^{-j\phi_1} \\ \vdots \\ a_{N-1} e^{-j\phi_{N-1}} \end{bmatrix} \quad (3)$$

where $\mathbf{v}^\dagger$ is the complex multiplier vector, r is the final received signal, a_n and ϕ_n are the gain and phase shift of the signal from the n th receive antenna. The complex coefficients $\mathbf{v}$ are typically set by a MIMO algorithm which is chosen according to a variety of criteria [3].

III. VECTOR DEMODULATOR ARCHITECTURE AND ALGORITHM

In this section a mixer-bank-based vector demodulator is introduced, and the $\Sigma\Delta$ approximation algorithm is reviewed.

A. Oversampling and Coarse Quantization-based Vector Demodulator

In one possible implementation of the vector demodulator, two VGAs could independently scale in-phase (I) and quadrature (Q) versions of the incoming signal, which together represent a complex phase-shift [4]. This system architecture, which is shown in Fig. 1, seems simple, but an RF VGA based vector demodulator is difficult to implement in practice because the VGAs must be sufficiently linear and have high-resolution in the region of interest to

achieve fine-resolution gain and phase shift control. The resolution requirement of the VGAs could be reduced if the target complex coefficient v_i were reconstructed from many LO phases. As the number of LO phases increases, at some point the gain resolution requirement reduces to a binary decision, which is inherently linear.

An example of the proposed vector demodulator with $M=4$, where M is the number of LO phases it employs, is shown in Fig. 2. The LO signals are denoted by their relative delays in degrees. The example in Fig. 2 oversamples by a factor of two in the phase domain. The gain resolution of the VGAs reduces to a ± 1 decision, which can be implemented by a simple cross-coupled CMOS switch quad. These switches have been moved to the output of the mixers to reduce the number of mixers by a factor of two. The binary coefficient b_m that sets each of these switches is generated by a digital algorithm and passed into the analog domain to control the vector demodulator. The addition at the output of the switches can be done by current combining.

Because the proposed vector demodulator shifts design complexity from the analog circuits to the digital algorithm, its analog components can be kept simple. The algorithms can be implemented with digital circuits, so it is possible to minimize the system's sensitivity to process, supply voltage, and temperature (PVT) variations.

The LO phases can be generated by a ring-type voltage-controlled oscillator (VCO) embedded in a wideband phase-locked loop (PLL) to clean up phase noise. Alternately, the clock phases can be derived from a single clock phase via a delay-locked loop (DLL), as shown in Fig. 2.

The complex constellation that can be approximated in an implementation with $M = 8$ bases is shown in Fig. 3. The bases are highlighted with stars, and are located on the unit circle. The vector demodulator can approximate any gain or phase-shift within the unit circle with small residual error.

B. $\Sigma\Delta$ Approximation Algorithm

Approximating a complex multiplication by interpolating among many different phases is equivalent to solving the problem

$$\min_{b_m} \left| v_i - \sum_{m=0}^{M-1} b_m e^{j \frac{\pi m}{M}} \right|^2 \quad (4)$$

where v_i is the target complex coefficient for the i th receive antenna to be approximated and b_m the weight associated with the m th LO phase, or basis. This is known to be an NP -hard problem, and many suboptimal techniques exist to search for the best approximation. Among these techniques are the stochastic approximation method [5] and the $\Sigma\Delta$ approximation method [6]. The stochastic approximation method does not require uniform LO phase spacing, but requires a different linear amplification after

Fig. 1. VGA based vector demodulator.

Fig. 2. Direct conversion implementation of the vector demodulator for the case $M=4$.

each mixer. The $\Sigma\Delta$ approximation does not require different amplification along each down-conversion path, but requires the LO phases be uniformly distributed between 0 and π . Because any multi-phase LO generation circuit will likely generate clock signals that are uniformly spaced in the phase domain, we focus on the $\Sigma\Delta$ approximation method.

In $\Sigma\Delta$ ADCs, the quantizer is embedded in a feedback loop in such a way that quantization errors in each cycle will be compensated for in the following cycle. By oversampling at a high rate, this error compensation between cycles allows the ADC to produce high resolution samples even when utilizing a low resolution quantizer. The state update equations for a $\Sigma\Delta$ ADC are:

$$\begin{aligned} e[n] &= y[n] + e[n-1] - Q(y[n] + e[n-1]) \\ \hat{y}[n] &= Q(y[n] + e[n-1]) \end{aligned} \quad (5)$$

where $y[n]$ is the unquantized input signal, $e[n]$ is the quantization error at the n th time-step, and $\hat{y}[n]$ is the quantized version of the input signal.

Similar to the $\Sigma\Delta$ ADC, when approximating any desired complex gain, it is possible to achieve high accuracy by oversampling in the phase domain, even when utilizing 1-bit multipliers. To initialize the proposed algorithm, the

Fig. 3. Vector constellation from the $\Sigma\Delta$ algorithm for $M=8$.

following variables are set:

$$\begin{aligned} y[m] &= \langle v_i, \phi_m \rangle \\ &= \text{Re}\{v_i\}\text{Re}\{\phi_m\} + \text{Im}\{v_i\}\text{Im}\{\phi_m\} \\ e[-1] &= 0 \end{aligned} \quad (6)$$

Here, we define $y[m]$ as the projection of the target complex coefficient onto the m th basis. The term e , which is the phase “quantization” error, is initialized to 0. The state update equations for the algorithm are [6]

$$\begin{aligned} e[m] &= y[m] + e[m-1] - Q_m(y[m] + e[m-1]) \\ \hat{y}[m] &= Q_m(y[m] + e[m-1]) \end{aligned} \quad (7)$$

where $e[m]$ is the error on the m th basis, Q_m is the quantization of its argument to ± 1 . The term $\hat{y}[m]$ is the approximation of $y[m]$ in the ϕ_m direction. Note the similarity between (5) and (7).

A reconstructed approximation to the target complex gain can be obtained by adding the approximations $\hat{y}[m]$, followed by normalization. This can be written as

$$\hat{v}_i = \frac{2}{M} \sum_{m=0}^{M-1} \hat{y}[m] \phi_m = \frac{2}{M} \sum_{m=0}^{M-1} \hat{y}[m] e^{j\frac{\pi m}{M}} \quad (8)$$

where $\hat{v}_i$ is the final reconstructed approximation to the desired complex gain. It also follows that $\phi_m = e^{j\frac{\pi m}{M}}$ because the $\Sigma\Delta$ algorithm requires that the bases be uniformly distributed between 0 and π . In the proposed vector demodulator architecture, the reconstruction in (8) happens in the analog domain. The $\Sigma\Delta$ algorithm runs in the digital part of the proposed receiver architecture, shown in the bottom half of Fig. 2. The outputs of the algorithm, $b_m = \hat{y}[m]$, cross over into the analog domain and control the vector demodulator. The reconstruction along the different bases ϕ_m in (8) are facilitated by the M different LO phases.

C. Comparison with existing architectures

The complex coefficient constellations of vector demodulators implemented with RF VGAs are limited in

Fig. 4. MSE vs. number of bases (without mixer mismatch).

TABLE I
COMPLEXITY FOR DIFFERENT BEAMFORMING APPROACHES.

	LO Phases Required	Number of Mixers	Number of ADCs
Digital Approach	2	$2N$	$2N$
Proposed Approach	M	MN	2

theory by the linearity of the VGAs. While the signal combining between the in-phase and quadrature VGAs may be linear, the VGAs themselves are not. In other words, both the linear portion and the nonlinear portion of the phase shifting algorithm in a RF VGA based vector demodulator are done in the analog domain.

In contrast, in the proposed approach the nonlinear portion of the oversampling algorithm is done in the digital domain where the computation is well-controlled, while the linear signal combination portion is done in the analog domain. As a result, the proposed architecture is inherently more linear than the RF VGA approach.

While increasing M without bound decreases the MSE to an arbitrarily low value, it also increases the hardware costs significantly. The hardware complexity for both the digital and proposed beamforming approaches are shown in Table I, where N is the number of receiving antennas. A typical state of the art mixer might consume between 2mW and 5mW, while a current 100MB/s 10b ADC might consume between 10mW and 20mW. Note that while the proposed approach requires the generation of more LO phases, it also relaxes the ADC requirements. We choose $M = 8$ to be a reasonable compromise between approximation accuracy and hardware complexity.

IV. SIMULATION RESULTS

In this section behavioral simulations of the proposed vector demodulator are presented. The accuracy of the vector demodulator with no analog impairments is discussed first to demonstrate the theoretical performance of the architecture. Analog impairments are then added, and the effects on accuracy are explained.

Fig. 5. MSE vs. phase mismatch: (a) $E[\text{MSE}(\sigma_\theta)]$, (b) standard deviation of $\text{MSE}(\sigma_\theta)$.

A. $\Sigma\Delta$ Algorithm Performance

Although the $\Sigma\Delta$ algorithm runs in the digital domain, the reconstruction of each gain vector will occur in the analog domain by the configuration of M complex mixer outputs. Therefore, the approximation accuracy of the algorithm is subject to the random phase mismatches between the mixers. This was taken into account in simulation by perturbing the phases of the reconstruction bases. The mean squared error (MSE) is denoted as $\text{MSE}(\sigma_\theta)$, where σ_θ is the standard deviations of the phase mismatches between the mixers.

Before phase mismatch was added, the accuracy of the algorithm was determined by using it to approximate 10,000 target complex coefficients. The MSE as a function of M , the number of bases, was then calculated. Each target complex coefficient has the form

$$v_i = Ae^{j2\pi B} \quad (9)$$

where both A and B are independent random variables uniformly distributed between 0 and 1. The MSE from the $\Sigma\Delta$ algorithm as well as the results of a brute force search of the best complex coefficient approximation are shown in Fig. 4. The brute force search is predictably more accurate than the $\Sigma\Delta$ algorithm, but we found that it is prohibitively complex for $M \geq 8$. As M increases, the MSE of the $\Sigma\Delta$ algorithm decreases in proportion to M^2 . With a large enough M , the error can be made arbitrarily small. This is consistent with [6], a theoretical study that states that the gain approximation error decreases asymptotically as M^2 , so long as the target to be approximated is bounded.

B. Sensitivity to Analog Impairments

The sensitivity of the approximation MSE to phase errors in the mixers is displayed in Fig. 5(a) and Fig. 5(b), which show $E[\text{MSE}(\sigma_\theta)]$ and the standard deviation of $\text{MSE}(\sigma_\theta)$ respectively. Because phase mismatch is zero-mean, $E[\text{MSE}(\sigma_\theta)]$ is not sensitive to small phase and magnitude variations in the mixers. The approximation error of

the $\Sigma\Delta$ algorithm also decreases in proportion to M^2 , as expected. It can be seen from Fig. 5(b) that the MSE error due to phase mismatch is not a strong function of M . This is because mismatches between the mixers are uncorrelated and tend to be averaged out during reconstruction, causing an overall decrease in system sensitivity to mismatch. The additional MSE in the approximation due to mismatch is a small fraction of the base MSE expected from the algorithm for reasonable values of σ_θ .

V. CONCLUSION

In this paper mixed signal MIMO beamforming architecture based on oversampling and coarse quantization was presented. This paper has demonstrated that by shifting system complexity to the algorithmic level, which can be implemented with digital circuits, one can avoid designing complex analog circuits, which are sensitive to PVT variations. Because the proposed architecture employs multiple versions of simple analog circuits, it is inherently linear, relatively insensitive to component mismatch, and can simultaneously produce accurate variable gains and phase-shifts.

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