A Self-steering I/Q Receiver Array in 45-nm CMOS SOI

Arpit K. Gupta and James F. Buckwalter
Department of Electrical and Computer Engineering, University of California, San Diego,
9500 Gilman Dr., La Jolla, CA - 92093

Abstract — **A novel I/Q receiver array is demonstrated that adapts phase shifts in each receive channel to point a receive beam toward an incident RF signal. The measured array operates at 8.1 GHz and covers steering angles of +/-35 degrees for a four element array. Additionally, the receiver incorporates an I/Q down-converter and demodulates 64-QAM with EVM less than 4%. The chip is fabricated in 45 nm CMOS SOI process and occupies an area of $3.45\,mm^2$ while consuming 143 mW dc power.**

Index Terms — **Retrodirective arrays, adaptive phased-arrays, self-steering, coupled-phased locked loops, X-band**

I. Introduction

Phased array transmitters and receivers provide high signal-to-noise ratio (SNR) in microwave and millimeter-wave communication links. As higher directivity is introduced between transmitter and receiver, an undesirable consequence is the need to accurately steer the desired beam to maximize SNR. In particular, point-to-point E-band (70/80 GHz) links have narrow beamwidths (<2 degrees) and require skilled labor to align high-gain transmit and receive antennas. A phased-array replacement should introduce a self-steering mechanism to enable full-duplex communication [1], [2]. Retrodirective arrays also require self beam-steering where the receive beam is steered in the direction of the interrogating signal. Full-duplex communication is established by receiving from and transmitting to the same direction. Fig. 1 illustrates the concept of a self-steering phased array receiver. The front-end downconverts each channel to a self beamsteering IC that forms a beam in the direction of the desired RF signal source. In other words, the self-steering beamformer provides the maximum receiver SNR irrespective of the direction of the incident signal. Ideally, the adaptation mechanism can steer over a complete $\pm 90^o$ range of incident RF angles.

The solution proposed here utilizes a coupled oscillator array and coupled-phased locked loop to adaptively steer to the incident direction of a desired signal. The receiver is also capable of demodulating I and Q data paths. The receiver circuit presented in this work is demonstrated with four elements but can be scaled to larger arrays. It operates for an RF input at X-band (8 - 12 GHz) and could be used in conjunction with a millimeter-wave front-end as shown in Fig. 1. Section II describes the operation of the adaptation circuitry and discusses the design issues. Section III describes the implemented circuitry. Measurement results are presented in Section IV to demonstrate maximum SNR with respect to different incident RF beam angles and operation with a 64-QAM constellation at 60 Mb/s with an EVM under 4%.

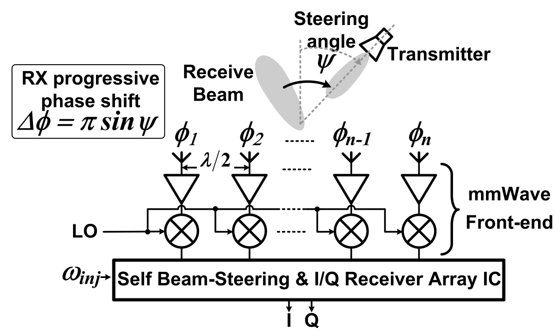

Fig. 1. Self-steering beamformer circuitry for receiver arrays. The receiver forms a beam in the direction of the interrogating signal.

II. Self-steering System Concept

A progressive phase shift $\Delta\phi$ between neighboring receive channels forms a beam toward a signal incident from an angle ψ as shown in Fig. 1. For $\lambda/2$ antenna spacing, $\Delta\phi = \pi \sin\psi$. The circuitry proposed here generates the progressive phase shift $\Delta\phi$ across the receiver elements by changing the phase of individual LOs in a coupled oscillator array (COA). To adaptively form a beam in a desired direction, a feedback mechanism is introduced that adjusts the phase of each channel and is realized through a coupled phased-locked loop (CPLL). The COA and CPLL operate with joint dynamics that are explained briefly here.

Fig. 2 illustrates two weakly coupled-oscillators where ϵ is the coupling magnitude, θ_i, ω_i and Q are respectively the instantaneous phase, natural frequency and quality factor of the i_{th} oscillator. For $\lambda/2$ antenna spacing, $kd = \pi$, where k and d are the phase constant and antenna spacing. When the natural frequency of the oscillators is detuned (i.e. $\omega_i \neq \omega_{i-1}$) within their locking range, the oscillators lock to the same frequency with a phase difference.

However, the stable oscillator phase difference $\triangle\theta$ is limited to $\pm 90^o$ which corresponds to a steering angle range of $\pm 30^o$ as shown in Fig. 2 [3]. To increase the steering range, a frequency multiplication M increases the

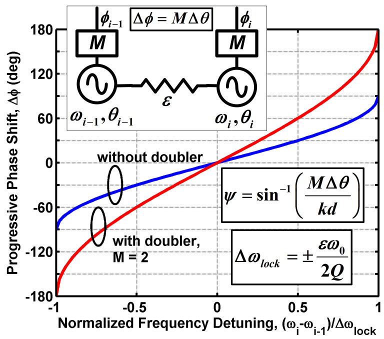

Fig. 2. Coupled-oscillator dynamics for beam-scanning.

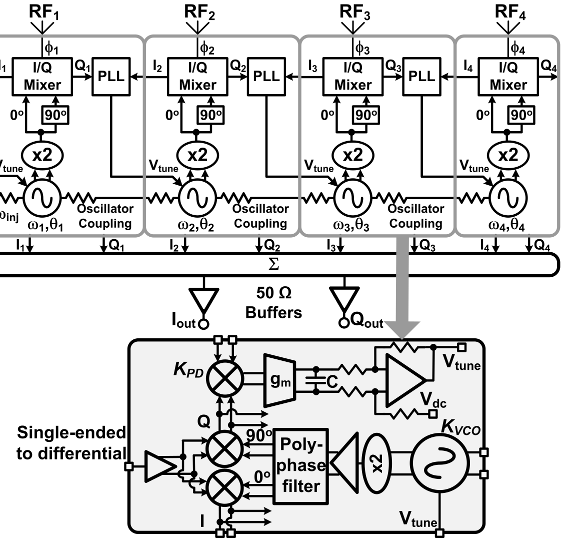

Fig. 3. System block diagram of a 4-channel self-steering beamformer circuit using coupled oscillators and coupled phased-locked loops.

phase difference $\Delta\phi$ and therefore the scanning range [4]. For instance, the RF phase difference $\Delta\phi$ extends to $\pm 180^o$ using frequency doublers ($M = 2$) and increases the steering range ψ to $\pm 90^o$.

In a larger COA, constant progressive phase shift is generated by setting the natural frequency of the interior oscillators to the same frequency ($\omega_i = \omega_o$) and detuning the outermost oscillators [5]. However, precise control of the ω_i is difficult due to circuit mismatches. Discrete COAs for transmit systems have demonstrated a limited constant phase progression from -20^o to $+50^o$ (or steering range equal to -6.5^o to $+16^o$) [6]. Demonstrating much larger steering ranges through integrated circuit techniques is one goal of the circuitry proposed here.

To adaptively steer the array and maximize the SNR, a CPLL is used to set the natural frequency - and hence phase - of each LO relative to an error signal measured from its neighbor. The oscillators in each channel remain locked together through the COA. Previous work on retrodirective arrays suggested open-loop operation using external variable phase shifters, which require careful calibration [1]. The CPLL provides the proper phase shifts while compensating for mismatch of each receiver channel.

Fig. 3 illustrates the combined COA and CPLL for oscillator control. The phase dynamics of each oscillator in the presence of the COA and the CPLL are described by (1), where $i = 2, 3$, A and ω_o are the oscillator amplitude and free-running frequency, K_{vco} is the oscillator tuning coefficient, K_{pd} is the phase detector gain, and $f(t)$ is the impulse response of the loop filter (integrator) and is equal to $f(t) = \frac{g_m}{C}u(t)$ where $u(t)$ is the unit step function. The 90 degree excess phase shift in $(i-1)th$ oscillator path is added to lock each oscillator in phase since the loop forces the phase difference between the input signals of the phase detector to be 90^o irrespective of the phase difference between the incident RF signals. Under steady-state locked conditions, the dynamics produce $V_{PD,i} = 0$ and $\frac{d\theta_i}{dt} = \omega_{inj}$. If the phase detector is driven by the quadrature and in-phase signals of the neighboring RX paths, the steady state phase shift between neighboring oscillators is

$$2\Delta\theta = \Delta\phi \tag{2}$$

where the factor of two arises from the use of a frequency doubler. Therefore, the CPLL produces the desired progressive phase shift across the phased array aperture. From (2), a stable phase difference of $\pm 90^o$ between coupled oscillators produces a $\pm 180^o$ progressive phase shift. Assuming $\lambda/2$ antenna spacing, this corresponds to $\pm 90^o$ steering range across the aperture.

III. Circuit Implementation

Fig. 3 shows the block diagram of the implemented self-steering I/Q receiver array. Four *LC* oscillators are arranged with nearest neighbor resistive coupling elements. An NMOS biased in the triode region allows varying the coupling strength from $\epsilon = 0.1$ to 0.5. The oscillator natural frequency is designed to be 5 GHz or half the RF frequency at 10 GHz. NMOS varactors are used in the *LC* tank as the tuning elements. The simulated oscillator tuning range is from 4.25 GHz to 5.4 GHz allowing

$$\begin{aligned}
\frac{d\theta_1}{dt} &= \omega_{o,1} + \frac{\epsilon\omega_{o,1}}{2Q}\left[\frac{A_{inj}}{A}\sin(\theta_{inj}-\theta_1) + \sin(\theta_2-\theta_1)\right] \\
\frac{d\theta_i}{dt} &= \omega_{o,i} + \frac{\epsilon\omega_{o,i}}{2Q}\left[\sin(\theta_{i-1}-\theta_i) + \sin(\theta_{i+1}-\theta_i)\right] + 2K_{vco}K_{pd}\cos\{(2\theta_{i-1}-\phi_{i-1}+90^o)-(2\theta_i-\phi_i)\} * f(t) \\
\frac{d\theta_4}{dt} &= \omega_{o,4} + \frac{\epsilon\omega_{o,4}}{2Q}\left[\sin(\theta_3-\theta_4)\right] + 2K_{vco}K_{pd}\cos\{(2\theta_3-\phi_3+90^o)-(2\theta_4-\phi_4)\} * f(t)
\end{aligned} \tag{1}$$

coverage of RF inputs between 8.5 and 10.8 GHz. The doubler is loaded by a parallel LC tank with $2f_{osc}$ as the center resonant frequency. The bandpass behavior of the LC tank filters higher order harmonics generated in the doubler. The doubler output is amplified and a single stage poly-phase filter generates the differential I/Q phases of the mixer LO.

At the front-end, an active balun converts the single-ended RF signal to a differential signal. Gilbert-cell mixers down-convert the signal from the RF to IF band. Gilbert-cell mixers exhibit low loss and noise figure at the expense of power consumption. To reduce the effect of parasitic capacitance and routing losses, a low IF frequency equal to 100 MHz is chosen in this design. The CPLL consists of a Gilbert-cell mixer as phase detector, gm-C integrator and op-amp in feedback as the summer. The phase detector is driven by the down-converted in-phase and quadrature IF signals of neighboring elements. Off-chip 10 nF capacitors are used for the gm-C integrator to improve the loop stability. Common-mode feedback circuit sets the output DC voltages at the integrator nodes. A conventional RC compensated two-stage op-amp in resistive feedback acts as the voltage summer. The op-amp unity gain BW is designed to be much higher than PLL loop BW (500 KHz). The in-phase and quadrature down-converted IF signals from each channel are combined using two separate combiners. The combined IF signals are buffered to an off-chip 50 ohm load where the demodulated signal can be measured.

IV. Measurements

The 4-element self beamsteering chip is fabricated in 45-nm SOI CMOS and is shown in Fig. 4. The chip area is 2.3mm by 1.5mm. For measurement, the chip is mounted on a PCB using conductive bond. All the pads are wire-bonded using standard 1 mil gold bond-wires to a Rogers 4000 substrate. The circuit board uses ground-CPW lines for high isolation between RF channels. The small 8 mils thickness of the substrate allows low inductive transition to the RF pads. The circuit consumes 143 mW power or 34 mW per channel.

The output of an Agilent E8257D analog signal generator is split four ways and four mechanical phase shifters are inserted in RF paths to emulate an RF beam coming from an arbitrary direction. An Agilent N5181A analog signal generator and 180 degree hybrid coupler injects a -3 dBm differential signal into the COA. An Agilent E4448A spectrum analyzer and DSO oscilloscope monitor the output IF signals in frequency and time domain.

The measured I/Q IF power is plotted in Fig. 5 with respect to the RF progressive phase at 8.1 GHz RF frequency and 4 GHz injected signal frequency. The 4-

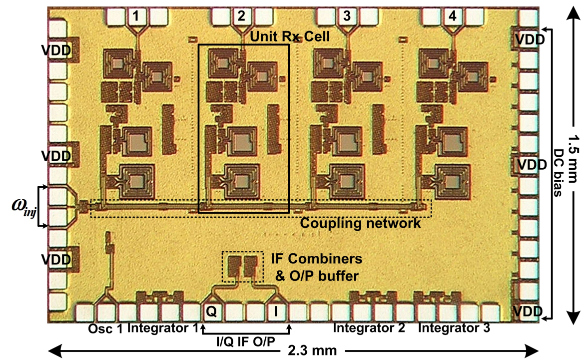

Fig. 4. Chip microphotograph.

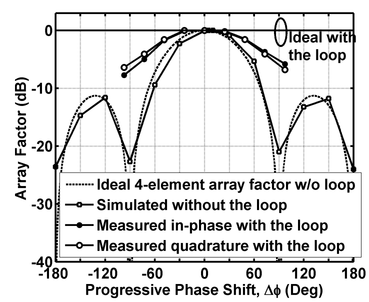

Fig. 5. Measured 4-element array factor and beam steering response.

element array factor is shown in Fig. 5 for a broadside beam in the absence of any self-steering.

When the CPLL is operating, the circuit ideally steers toward the incident RF beam and should produce no degradation in the SNR. The array achieves $\pm 100^o$ linear phase range corresponding with a $\pm 35^o$ steering range for half-wavelength antenna spacing. Notably, the amplitude is degraded as the COA is directed toward larger phase shifts. This amplitude degradation is related to changes in the oscillator amplitude with large array phase shifts. Nonetheless, the steering produces an 18 dB power improvement at the array factor nulls ($\pm 90^o$) over a scheme without adaptation.

The 3-dB IF bandwidth of the implemented array is 240 MHz as shown in Fig. 6. To demonstrate the demodulation performance of the array, an M-QAM modulated RF signal is provided to the system. The modulated RF signal is generated by upconverting a modulated IF signal using Agilent N5182A vector signal generator and an external mixer. Digital demodulation utility of Agilent E4448A spectrum analyzer is used to measure the error-vector-magnitude (EVM). Fig. 7 shows the measured output constellation and I/Q eye diagrams for 64-QAM at 10 MS/s symbol rate (or 60 Mbps bit rate) with progressive phase shift of -73 degrees. Input RF power is -16 dBm per channel. The measured EVM is 3.4%. The array could not be measured beyond 10 MS/s symbol rate due to the IF

BW limitation of spectrum analyzer. However, the large IF bandwidth indicates operation to higher data rates without significant EVM degradation. The modulated input RF and output IF spectra are plotted in Fig. 8 for 64-QAM at 10 MS/s symbol rate. The output SNR degrades by 12 dB due to receiver noise. The spectral re-growth seen in the input RF spectrum is due to the mixer and signal generator non-linearity.

Table I compares the implemented monolithic array with prior discrete demonstrations of retrodirective arrays. The array achieves the highest adaptive progressive phase range with 4 elements at highest operational frequency.

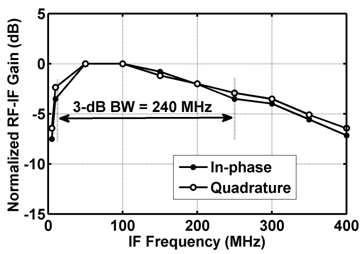

Fig. 6. Measured normalized conversion gain (RF-IF) at fixed $f_{osc} = 4$ GHz.

TABLE I
PERFORMANCE SUMMARY AND COMPARISON

	[1]	[6]	**This Work**
Integrated	No	No	Yes - CMOS 45nm SOI
Elements	2	3	4
RF Frequency (GHz)	1.425	2.68 - 2.72	7.4 - 9.4
Stable Progressive Phase, $\Delta\phi$	$\pm 156^o$ (Open-loop)	-20^o to $+50^o$	$\pm 100^o$ (4-elements) $\pm 170^o$ (3-elements)
Power Consumption (mW)	–	–	143 (including combiners & 50 Ω buffers)

V. Conclusion

A self-steering beamforming receiver has been proposed and fabricated using a 45-nm CMOS SOI process. The adaptive array is capable of steering over a range of more than ± 35 degrees and can produce stable progressive phase shifts of ± 100 degrees across the receive array. The demodulated signal in the array exhibits EVM under 4% for 64-QAM.

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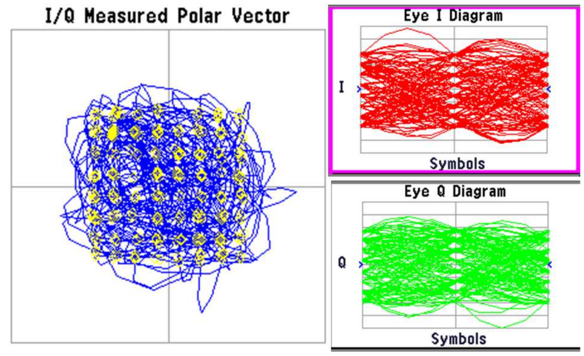

Fig. 7. Measured output constellation and eye diagrams for 64-QAM at 10 MS/s symbol rate for -16 dBm RF input. EVM is 3.4%.

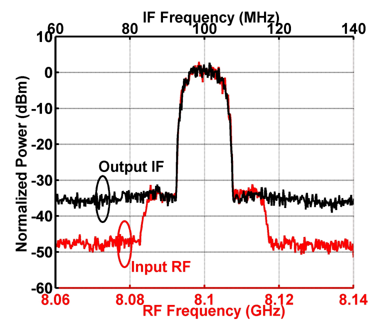

Fig. 8. Measured modulated spectrum with 64-QAM at 10 MS/s symbol rate.

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