Retro-Transceiver Array Using Monopole Antennas

S. Karode and V. Fusco

***Abstract*—In this paper, we show how a self-tracking antenna array constructed using $\lambda/4$ monopoles can be constructed, which is capable of receiving with gain over an entire 360° azimuthal cut. It is also shown how the self-tacking receive unit can be used in conjunction with a self-phased transmitter so that self-steered spatially selective receive and transmit functions can be formed simultaneously. The resulting array is capable of maintaining spatially selective receive and transmit functions to a roaming target without prior knowledge of its physical location.**

***Index Terms*—Monopole, retrodirective, self-steering.**

I. Introduction

Self-phased or adaptive antenna arrays may be defined as antennas in which each element is independently phased for coherent reception/transmission based on the information obtained from the received signals. Here, each element phases itself without *a priori* knowledge of its position relative to the source [1]. The objective being to achieve beam steering, which is automatic and always in the direction of arrival of the incident signal.

Two basic architectures for self-phased arrays have been reported, which are based on different techniques used for phase conjugation [2], [3]. Both are suitable for various communication applications such as direct satellite broadcast (DBS) [4], [5]. The receive antenna architecture presented here is different from the antennas mentioned above, which are repeaters and which can only be modified to produce conventional receive only functionality. That is, they can only form the receive beam maximum response at boresight to the array axis. Hence, the array presented here strives to achieve the advantages of an omnidirectional antenna [6], i.e., maximum response in all receive directions, with that of a steered antenna i.e., high gain in the direction of the incoming signal [7] without the need for auxiliary control hardware/software.

The retrodirective antenna arrays in [1]–[3] were designed for self-directed retransmission of an incoming signal. This paper presents a new architecture for the self-steering receive portion [8] of a retro-transceiver array. The self-steering receive-only section is different in its mode of operation from previously reported steered receiver antennas, e.g., [9]–[12], The work presented in this paper is the subject of a British patent application [8].

II. Two-Element Retroreceive Array Operation

A two element embodiment of the retroreceive antenna is shown in Fig. 1. Its operation is described below.

In Fig. 1(a), when an incident signal arrives at any angle other than boresight, a phase delay ϕ is introduced into the signal received at each element comprising the array. If both the elements are equidistant from the array center, then the signals received by these elements will bear a phase conjugate relationship with respect to each other.

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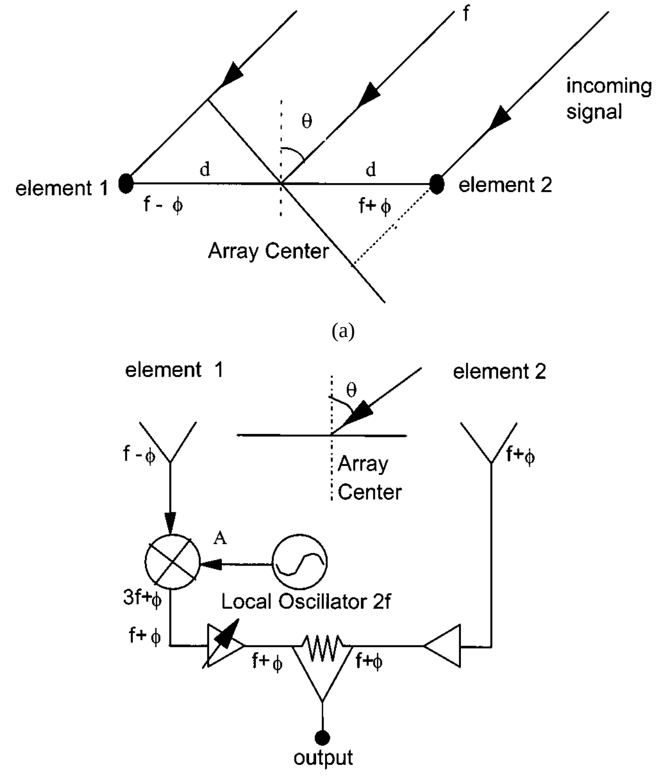

Fig. 1. Two-element embodiment of retroreceive antenna. (a) Incident wavefront. (b) Retroreceive structure.

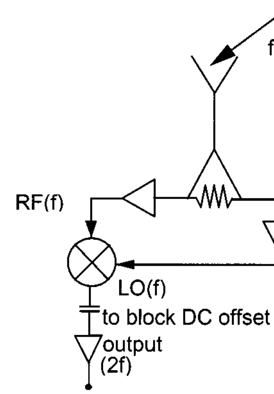

Fig. 2. Generation of reference signal.

Let the signals received by elements 1 and 2 in Fig. 1(b) be

$$e^{j(\omega t-\phi)} \text{ and } e^{j(\omega t+\phi)}, \text{ respectively} \tag{1}$$

where ω_t is the frequency in radians and ϕ is the phase difference with respect to the signal received at the array center with angle θ measured with respect to the array boresight.

The signal from element 1 is mixed with a reference signal at twice the frequency of the incoming signal. The output of the mixer placed at element 1 will consist of two basic components, i.e., the sum and difference product. Here, the difference product has the same frequency as that of the input to the mixer, but with conjugate phase relationship

$$e^{j(2\omega t-(\omega t-\phi)}, \text{ i.e., } \quad e^{j(\omega t+\phi)}. \tag{2}$$

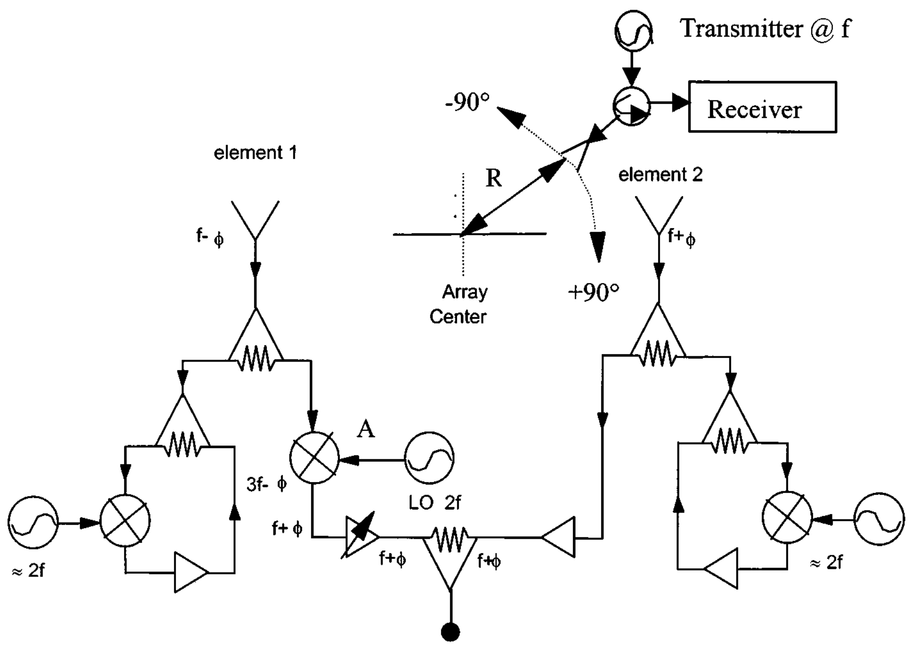

Fig. 3. Two-element retrodirective transceiver array.

The $e^{j(\omega t+\phi)}$ output of the mixer, after suitable amplification, and the signal from element 2 are added together using a power combiner to give an in-phase power-combined response for any angle of arrival of the incoming signal in the azimuthal plane.

Consider now the situation when the local oscillator (LO) has a relative phase shift with respect to the incoming signal. Let this relative phase difference be α_t.

Here, when the incident signal is at an angle θ_1, then the signals at the two elements relative to the array center (Fig. 1) will be for elements 1 and 2, respectively,

$$\omega t - \phi_1 \text{ and } \omega t + \phi_1 \tag{3}$$

where $\phi_1 = ((2\pi d)/\lambda)\cos\theta_1$.

Here, phase ϕ_1 is measured with reference to the phase of the incident signal at array phase center and d is the antenna separation.

On passing through the mixer, but before summation the signal, at element 1, will be

$$\omega t + \alpha_t + \phi_1 \text{ and for element 2 } \omega t + \phi_1. \tag{4}$$

Next, let the incident signal come from a different angle θ_2. Here, let the signals at elements 1 and 2 be, respectively,

$$\omega t - \phi_2 \text{ and } \omega t + \phi_2 \tag{5}$$

with LO relative phase shift included and before summation, the signals at the mixer output and element 2 are, respectively,

$$\omega t + \alpha_t + \phi_2 \text{ and } \omega t + \phi_2. \tag{6}$$

The phase changes occurring at these elements while shifting the angle of incident signal from θ_1 to θ_2 can be obtained by taking the difference of phases at these elements at positions θ_1 and θ_2 i.e., by subtracting (6) and (4). Thus, the phase change at the mixer output will be

$$(\omega t + \alpha_t + \phi_1) - (\omega t + \alpha_t + \phi_2) = \phi_1 - \phi_2 \tag{7}$$

and at element 2

$$(\omega t - \phi_2) - (\omega t - \phi_1) = \phi_1 - \phi_2. \tag{8}$$

Equations (7) and (8) show that, as long as the target is in motion, the relative phase differences of the received signals at elements 1 and 2 remain the same even after the angle of arrival of the incident signal has changed. Thus, the desired constant output response for all the azimuthal positions is maintained.

The reference signal used as an LO signal for the mixer can be generated by extraction from the signal received from the reference antenna placed at the array center (Fig. 2). The use of this signal ensures effective locking to the target signal even during fluctuations in its frequency. Here, the up-converted mixer output signal after suitable amplification is connected to the output, i.e., port A, in place of the $2f$ LO drive in Fig. 1(b). This signal $2f$ when mixed with the $f - \phi$ primary signal available at element 1 gives $f + \phi$ as before.

III. Retrodirective Transceiver Array

We now show the use of the retroreceive configuration in a self-steered transceiver (i.e., self-steering/self-tracking) system, Fig. 3. In the work presented in this paper, the element separation is $\lambda_0/2$ at 1 GHz. Both receive and retransmit function have concurrent self-tracking capability. Such a system could have applications in next-generation mobile communication applications where spatial division multiple access (SDMA) is important [12]. Previous to the work presented in this paper, either: 1) self-tracking retrodirective action, i.e., reception followed by automatic retransmission or 2) normal passive combining boresight maximum beam formation coverage by the receiver without the use of a supplementary pilot tone or additional electronics [4], [9]–[11] were the only possibilities when using self-tracking antenna technologies.

The transceiver array in this paper exhibits the capability of concurrent automatic self-steering of both transmit and receive polar patterns in the direction of the incoming signal. We use a conventional Pon architecture [3] for the retrodirective transmit section. Here, the retransmit frequency at 990 MHz is slightly offset from the receive frequency at 1 GHz ($\mathrm{LO} = 2$ GHz). This is done to enable receiver dis-

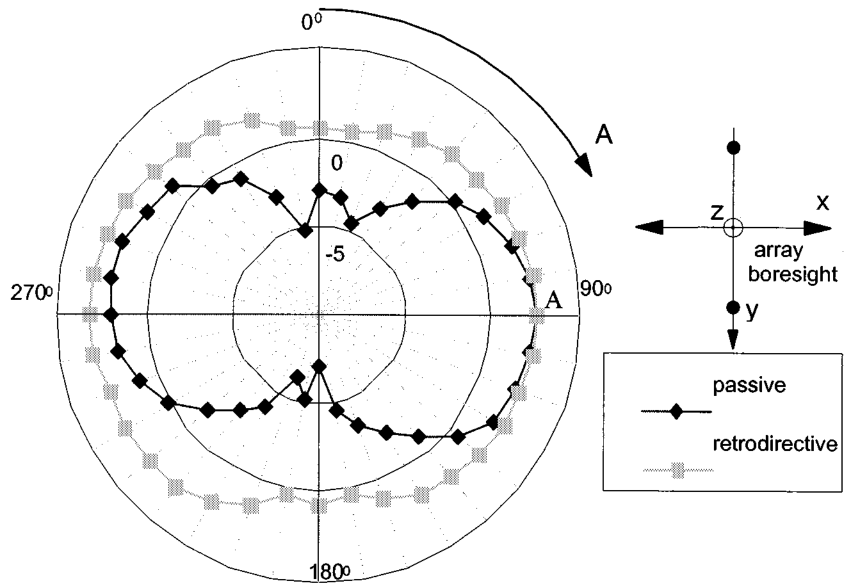

Fig. 4. Measured locus of normalized E-field maximum of the receive beam of the retrodirective transceiver array.

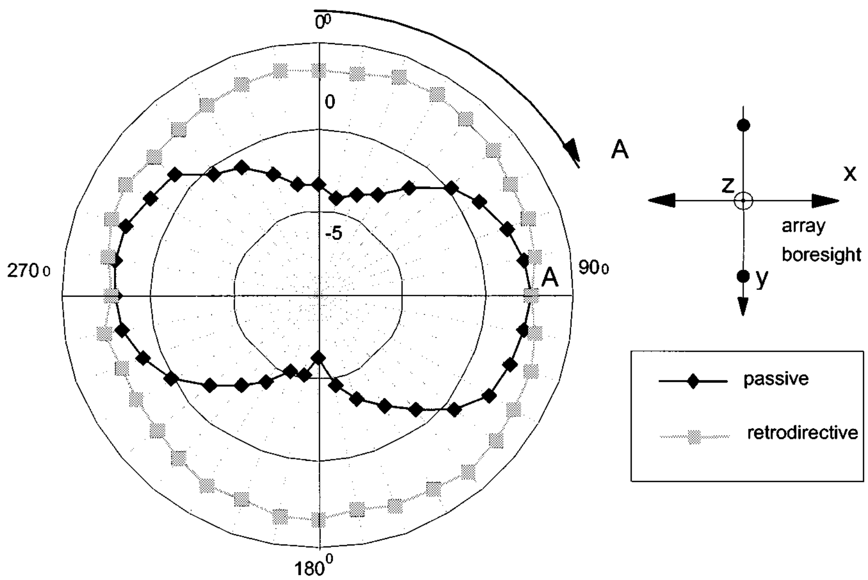

Fig. 5. Measured locus of normalized E-field maximum of the retransmit beam of the retrodirective transceiver array.

crimination in the monostatic measurement system (Fig. 3) between incoming and outgoing signals. This will lead to a small frequency squint in the return signal. The retroreceive configuration [see Fig. 1(b)] is used to form the self-steering receive section. The separation distance R in Fig. 3 is $10\lambda_0$. Figs. 4 and 5 show the receive and retransmit responses of a two-element retrodirective transceiver array, respectively. Monopole antennas have been used as the radiating elements in the array, each has a return loss of better than -25 dB at 1 GHz. The retrodirective traces in Figs. 3 and 4 are monostatic measurements (see Fig. 3), which show the locus of point A, i.e., the nominal boresight maximum response of the passive array as is self tracks, in receive Fig. 4 and retransmit Fig. 5 modes, respectively.

For reference, the radiation pattern of a two-element passive array constructed using the same dipole elements is also included in both Figs. 4 and 5. Measured results for the example discussed here show that the passive array provides approximately 3-dB beamwidth coverage of 60° in both transmit and receive modes. Figs. 4 and 5 show that the retrodirective transceiver array is able to provide (to within a 3-dB variation in signal) coverage in the entire azimuthal plane from 0° to 360° in both transmit and receive modes.

IV. Conclusions

This paper has demonstrated a novel self-steered receive-only antenna configuration, which was then incorporated into a compact transmit/receive unit that has self-steering capability on both transmit and receive functions simultaneously. Such a system has been shown to be capable of tracking with gain over 360° azimuthal coverage.

The approach of this paper has suggested one possible low-cost solution for systems that deploy SDMA as a means for achieving increased density of frequency reuse. The architecture developed in this paper could find applications in asset tracking and in next-generation mobile communications applications where conventional adaptive array techniques are either too expensive or have over-specified performance in mobile wireless applications where less demanding than fully adaptive beam-formed far-field radiation-pattern responses are required.

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Measurement of Two-Tone Transfer Characteristics of High-Power Amplifiers

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***Abstract*—In this paper, we present an accurate measurement method for acquiring the two-tone transfer characteristics of high-power amplifiers. The measurement setup and sequence are described. The measured amplitude and phase data of the two-tone fundamental, third-order intermodulation, and fifth-order intermodulation components versus input power level are also presented. The measured two-tone transfer characteristics are very useful for the design of a predistortion linearizer or for nonlinear model extraction for high-power amplifiers.**

***Index Terms*—AM–AM, AM–PM, high-power amplifier, memory effect, two-tone transfer characteristics.**

I. INTRODUCTION

The behavioral or mathematical model of power amplifiers has been studied extensively. A small class-A power amplifier has normally been treated with the assumption that is a memoryless (Taylor series representation of AM–AM characteristics only) or quasi-memoryless (complex representation of Taylor series with both AM–AM and AM–PM characteristics) system [1]–[5]. However, the characterization of a very high-power amplifier with an output power of over a few hundred watts has not been reported yet. Multistage high-power, class-AB or class-B amplifiers generally have a large memory effect and strong nonlinearity. The single-tone transfer characteristics based on AM–AM and AM–PM conversion cannot properly express the nonlinearity of these high-power amplifiers. Bosch *et al.* reported on a case where a predistortion linearized amplifier with improved AM–AM and AM–PM characteristics did not provide any enhancement on the two-tone intermodulation (IM) nonlinearity [6]. Therefore, more accurate two-tone characterization with phase information should be developed.

A method for measuring the relative phase of third-order intermodulation (IM3) compared to the phase of input signal has been presented by Suematsu *et al.* [7]. His method is based on the assumption that, in the weakly nonlinear region, the relative phase of IM3 is equal to the relative phase of fundamental. To verify the measured relative phase of IM3, they employed Volterra-series analysis based on the measured single-tone characteristics (AM–AM and AM–PM). Unfortunately, his assumption is not valid for most high RF power amplifiers. For a wide range of gate biases in GaAs MESFETs, the third-order Volterra-series coefficient of transcunductance (gm3) has reversed polarity to gm [8]–[10]. This means that the relative phase of IM3 may be 180° out of phase to the fundamental output. Moreover, he overlooked the internal device capacitance effect on the phase variation. At a high frequency, these capacitances may change the phases of fundamental and IM3. These phase variations can be easily verified through a harmonic-balance simulation with a large-signal model of MESFETs.

In this paper, we present a new accurate measurement technique for determining the two-tone transfer characteristics of high-power amplifiers. The relative phases of the harmonic terms of a very low-frequency amplifier are 0° or 180°. A small power GaAs MESFET amplifier at 750 kHz is used for the reference IM generator. For the measurement, the amplifier output is down-converted to the IF frequency and the relative phase is measured by comparing with the reference signal. A 500-W class-AB multistage power amplifier is used for the measurement. The measurement setup and sequence are described and the measured results are shown.

II. EXPERIMENT

A. Main Amplifier Under Test

A four-stage amplifier is built for Korea's wireless local loop (WLL) band of 2.37–2.4 GHz. Its final stage is built using four balanced 130-W amplifiers with Motorola's RF LDMOSFET MRF21120. It is a push–pull type with class-AB operation. The other three stages are arranged to drive the final stage amplifier. The peak output power at the 1-dB gain compression point is about 500 W and the overall gain is 44.5 dB. The operational average output power is 50 W with WCDMA signal input, which has a chip rate of 8.192 Mc/s. Fig. 1 depicts a line-up diagram of the main amplifier used for testing.

B. Measurement Setup

The measurement setup is shown in Fig. 2. This setup requires many measurement instruments: a two-tone signal generator, a vector network analyzer, a two-input power meter, and two spectrum analyzers. The reference IM generator is a small power MESFET of Hewlett-Packard's ATF21186 and is operated at a very low center frequency of 750 kHz. In the low frequency, the memory effect of the device

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