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Phase Error Effects on Distributed Transmit Beamforming for Wireless Communications

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***Abstract*—This paper investigates the impact of phase errors of beamforming networks on the performance of distributed transmit beamforming systems. Through multi-tone signal, wider band, models and the defined phase error percentage (PEP) of the beamforming networks, the distorted signal waveforms for different wider band occupying beamforming systems are presented. Furthermore, simulated bit error rates (BERs) are obtained to illustrate how the distributed array aperture sizes, the signal bandwidths, the PEPs, and the signal to noise ratios (SNRs) interact with each other. These studies are then used to provide some guidelines for wideband distributed transmit beamforming system design.**

***Keywords*—beamforming; bit error rate (BER); distributed transmit array; wideband; wireless communications**

I. Introduction

Alongside the benefits brought by the broadcast nature of wireless communications, there are weaknesses, such as wasted transmit energy and interference towards unwanted areas, reduced link budgets between communication nodes, and information leakage. Transmit beamforming, as a physical-layer communication means, utilizes multiple transmit antennas to form a high gain radiation beam towards a desired receiver, in such a fashion to alleviate the aforementioned weaknesses. Transmit beamforming is achieved by manipulating phases, and perhaps the magnitudes of the signals applied to the antenna array elements for sidelobe control, of the excitation signals at each transmit antenna, such that the differences of phase delays introduced by wireless propagation channels between each transmit antenna and the desired receiver can be compensated. This results in constructive signal combination at the receiver side, and destructive combination at other directions or locations.

Transmit beamforming can be carried out using a classical centralized antenna array, where all antenna elements are arranged in a regular pattern within a confined spatial region, normally in the order of tens of wavelengths. Alternatively, in wireless sensor networks sensor nodes can be scattered arbitrarily within a large area up to hundreds of meters and may need to collaboratively perform distributed beamforming in order to transmit commonly shared information back to a distant receiver node in an energy efficient way [1], [2].

The excitation weights used to enable transmit beamforming is normally generated with the help of beamforming networks at radio frequency (RF) stage, such as the simplest trombone line, the Butler Matrix [3], and the Rotman lenses [4], etc. For narrow frequency band beamforming, the beamforming network can be readily designed and the errors involved can be calibrated out. However, when transmitting signals occupying a wider bandwidth, e.g., in millimeter-wave communications, the design of the required beamforming networks can be complex due to the material dispersion and narrowband feature of some phase shifter networks. The imprecision of the frequency characteristics of the wideband beamforming network may have little impact in centralized transmit arrays. However, in distributed transmit arrays that may span an area with a diameter up to hundreds of metres, small imperfections in the beamforming networks may lead to a failure of information recovery at the desired receiver end.

In this paper we focus on the impact that phase shift errors have on distributed network performance for transmit beamforming. In Section II a model for the transmit beamforming system is established, and the effect of phase shift errors on received signal waveforms is illustrated. In Section III the distortion of these signals, using multi-tone signals as examples, are evaluated via extensive bit error rate (BER) simulations, through which some system design guidelines can be obtained. Finally, conclusions are drawn in Section IV.

II. Transmit Beamforming

Beamforming is a technique that is able to combine identical copies of signals transmitted by different transmit antennas constructively or, in other words, in-phase at the desired receiver direction or location. For the narrow frequency band signal transmissions, we use a one-tone signal $S_1(t)$ at the frequency f_c in (1) as an example,

$$S_1(t) = \exp(j2\pi f_c t). \tag{1}$$

Since only phase errors are investigated in this paper, the magnitudes of all signals are set to be unity. When identical copies of S_1 are radiated by N transmit antennas each with an excitation weight W_n (n = 1, 2, …, N), the signal detected by the desired receiver can be expressed as

$$R_1 = \sum_{n=1}^{N} W_n S_1\left(t - l_n / c\right) = \sum_{n=1}^{N} W_n \exp\left[j2\pi f_c \left(t - l_n / c\right)\right], \quad (2)$$

where l_n refers to the path length between the n^{th} transmit antenna and the receiver, and c denotes the speed of light. In order to enable constructive signal combination, the transmit antenna excitation weights W_n have to be designed as

$$W_n = W_1 \exp[j2\pi f_c(l_n - l_1)/c]. \quad (3)$$

When transmitted signals occupying a finite bandwidth are considered, the excitation weights W_n have to be wider band, satisfying

$$W_n(f) = W_1(f)\exp[j2\pi f(l_n - l_1)/c], \quad (4)$$

where the frequency f spans the entire signal bandwidth. Here the wireless propagation channel is assumed to be flat-fading within the signal bandwidth.

When non-ideal wider band weights $W_n(f)$ are used for beamforming, the signal waveforms could experience distortion at the receiver end. In this paper we introduce a parameter called '*Phase Error Percentage (PEP)*' over the signal bandwidth in order to quantify the effect that weight imperfection causes. *PEP* is defined as absolute phase errors versus ideal phase shifts at each frequency point, see (5). The function 'mod(x, 2π)' in (5) means taking the remainder when x is divided by 2π. In order to facilitate discussion in this paper it is assumed that the *PEP* is constant within the signal occupying bandwidth. It is noted that with the above definition a 100% *PEP* means the phase shifts are invariant with respect to frequency, which is the property of some coupler-based phase shifter types [5].

$$PEP(f) = \left| \frac{\mathrm{mod}\left\{ phase\left[W_n(f)\right] - phase\left[W_1(f)\right], 2\pi \right\}}{\mathrm{mod}\left[2\pi f\left(l_n - l_1\right)/c, 2\pi\right]} - 1 \right| \times 100\% \quad (5)$$

In order to visualize the impacts of the phase errors in the weights $W_n(f)$ have on received signal waveforms, here, multi-tone signals (M tones) with uniform frequency spacing Δf, like in orthogonal frequency-division multiplexing (OFDM), are constructed and transmitted by an N-element antenna array. The transmit array can be locally or distributely disposed. In order to facilitate simulation and results comparison the propagation path length differences $\Delta l = l_n - l_{n-1}$ are all set to be identical.

In Fig. 1(a) the normalized waveform of a 10-tone ($M = 10$) signal occupying 20 MHz (Δf = 2 MHz) is shown. After transmission through a 20-element array ($N = 20$) with a Δl of 0.1 m (centralized transmit array), it is found in simulated received signal waveforms in Fig. 1(b) and (c) that even if *PEP* is as high as 100% little signal waveform distortion can be observed. In simulations the excitation weights $W_n(f)$ was calculated using (4) subject to the designated *PEP*s.

However, while considering distributed transmit arrays where the path length differences Δl can be several meters, the

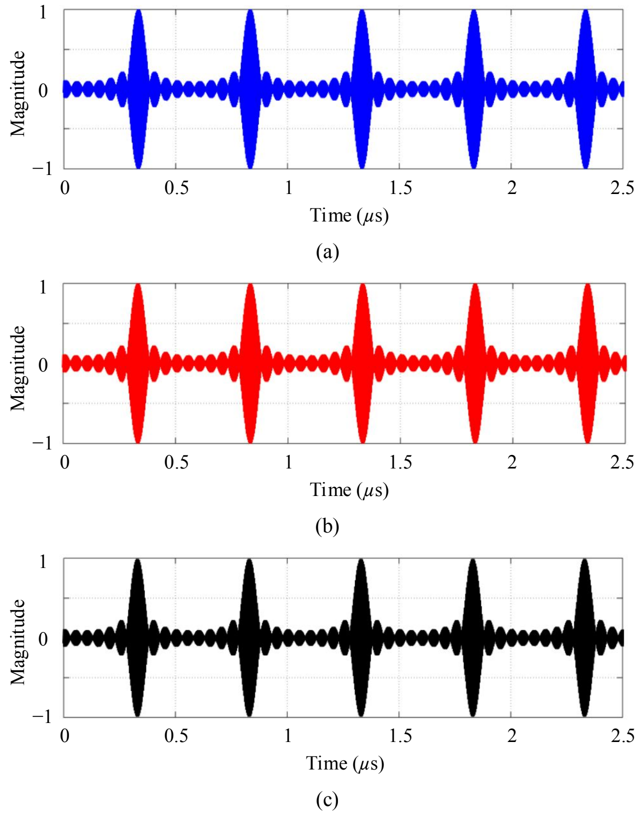

Fig. 1. Simulated wavefroms of (a) normalized transmited multi-tone signal when $M = 10$, $N = 20$, $\Delta f = 2$ MHz, and $\Delta l = 0.1$ m, (b) received signal when *PEP* = 20%, and (c) received signal when *PEP* = 100%.

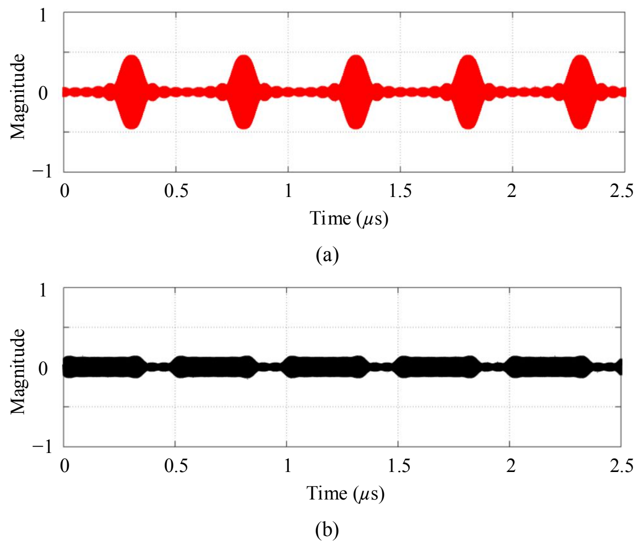

Fig. 2. Simulated wavefroms of (a) received signal when $M = 10$, $N = 20$, Δf = 2 MHz, $\Delta l = 5$ m, and *PEP* = 20%, and (b) received signal when $M = 10$, $N = 20$, $\Delta f = 2$ MHz, $\Delta l = 5$ m, and *PEP* = 100%.

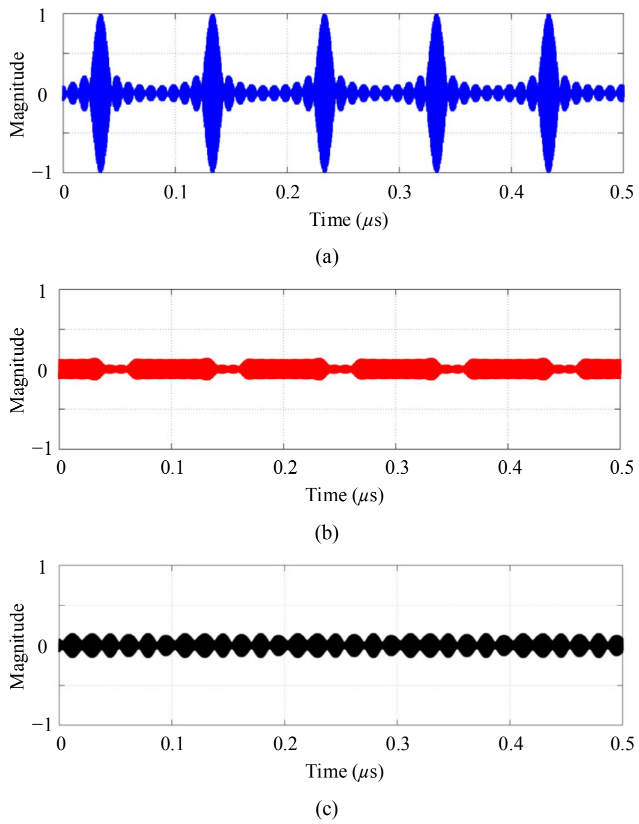

Fig. 3. Simulated wavefroms of (a) normalized transmited multi-tone signal when M = 10, N = 20, Δf = 10 MHz, and Δl = 5 m, (b) received signal when PEP = 20%, and (c) received signal when PEP = 100%.

received signal waveforms can experience severe distortion with the same amount of phase errors in the weights $W_n(f)$, see simulated received signal waveforms in Fig. 2 as examples when Δl is set to 5 m.

When the signal bandwidth is further widened (here we increase the Δf from 2 MHz to 10 MHz), the simulated received signal waveforms under different *PEP*s are plotted in Fig. 3(b) and (c). These waveforms are so distorted that it is hard to link them with the original transmitted signal copy in Fig. 3(a).

III. Bit Error Rate (BER) Simulation Results

In the last section it is shown through the simulated transmitted and received signal waveforms that the wider band distributed transmit beamforming is highly susceptible to the phase errors in array excitation weights. In order to provide guidelines for wider band transmit beamforming system design, e.g., the maximum acceptable phase errors subject to certain amount of signal quality degradation, or available wireless link budgets, the system BERs are obtained and presented in this section.

The BER simulations are conducted using the following prerequisites:

- Wider band signals are constructed using multiple tones (number M) with uniform frequency spacing Δf. Each tone is modulated with the same modulation schemes, e.g., QPSK and 16QAM. The data bits for each tone are random and independent. No data coding is applied.
- An N-element transmit array is considered. The wireless propagation path length differences Δl are assumed to be identical.
- Since multiple tones are evenly spread in frequency domain, the separation of them can be achieved using Fast Fourier Transform (FFT) modules, similar to an OFDM receiver. 64-point FFT is used in the simulation.
- Channel noise is assumed to be additive white Gaussian noise (AWGN) with zero mean.
- 10^{+7} random bits are transmitted in beamforming systems for each BER simulation, which allows BER down to 10^{-5} to be calculated.

In our simulation model, the aperture size of the distributed transmit array can be adjusted by both the number of array elements, i.e., N, and the path length differences Δl. As expected, larger aperture sizes, i.e., $N\times\Delta l$, contribute to higher BER values in wider band occupying beamforming systems, see simulation results in Fig. 4. Similarly, it is shown in Fig. 5 that the greater the signal bandwidths, i.e., $M\times\Delta f$, the higher BERs the receivers get when the signal to noise ratio (SNR) is fixed. As can be concluded from Fig. 4 and Fig. 5, when the PEP is fixed the performance of these beamforming systems is determined by the array aperture size and the signal frequency bandwidth, but not by the four individual parameters, i.e., N, Δl, M, and Δf. The BER curves under ideal beamforming conditions, i.e., PEP = 0, are also depicted for comparison in Fig. 4 and Fig. 5. They follow the classic QPSK BER-SNR relationships stated in [6], indicating that the multiple QPSK modulated tones are perfectly separated. For modulation types other than QPSK, similar results can be obtained but are omitted here due to the page limits.

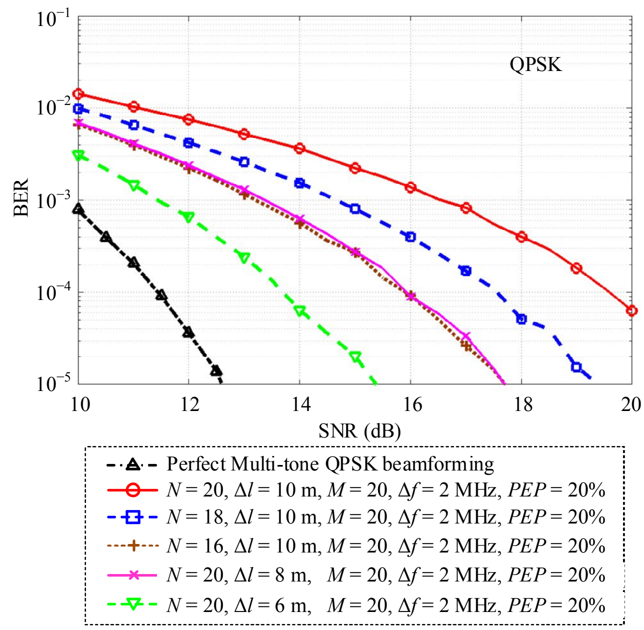

Fig. 4. Simulated BER versus SNR in distribured beamforming systems with various array aperture sizes. Each tone is QPSK modulated.

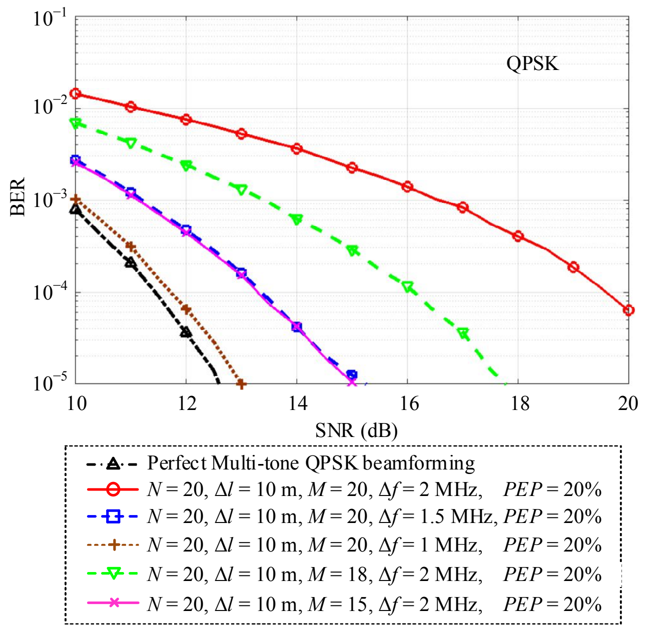

Fig. 5. Simulated BER versus SNR in distribured beamforming systems with various signal bandwidths. Each tone is QPSK modulated.

For a practical wider band occupying distributed beamforming system design, the array aperture size and the adopted signal bandwidth are normally fixed and known to the designers, thus graphs like examples shown in Fig. 6 and Fig. 7 can be useful. With the known aperture size, the signal bandwidth, the modulation type, and the targeting raw data BER, the *PEP* of the beamforming network and the required extra signal power can be directly linked. When system noise has constant power, the required extra signal power can be interpreted as the extra SNR, denoted as ΔSNR in Fig. 6 and Fig. 7. Take the plot in Fig. 6 as an example, when $N\times\Delta l = 160$ m, $M\times\Delta f = 40$ MHz, and the BER target is 10^{-3}, a *PEP* of 20%

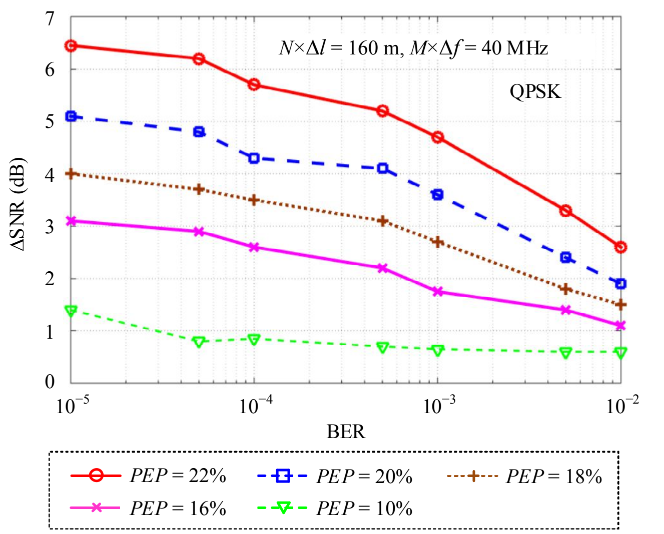

Fig. 6. Simulated extra SNRs required versus system BER targets for various PEPs of the beamforming networks. $N\times\Delta l = 160$ m, $M\times\Delta f = 40$ MHz, and each tone is QPSK modulated.

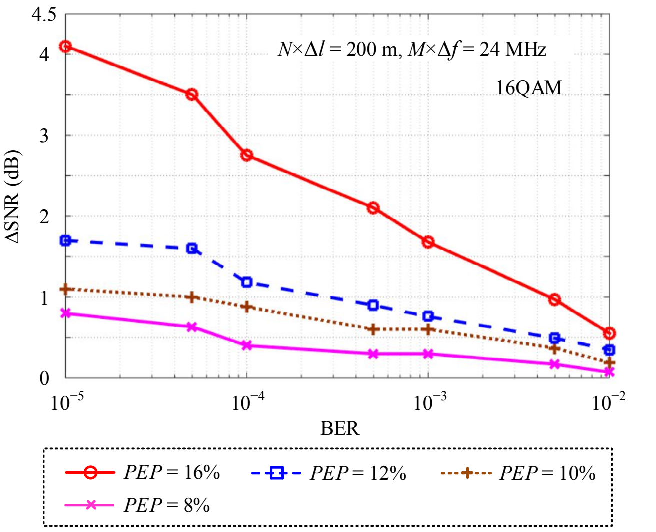

Fig. 7. Simulated extra SNRs required versus system BER targets for various PEPs of the beamforming networks. $N\times\Delta l = 200$ m, $M\times\Delta f = 24$ MHz, and each tone is 16QAM modulated.

indicates that an extra of 3.6 dB SNR is needed to achieve the same performance as in the ideal corresponding beamforming system. While from the other perspective, a link budget of 3.6 dB can only tolerate the *PEP* of the beamforming network up to 20%.

IV. Conclusions

It has been shown in this paper that in distributed transmit arrays the performance of wider band beamforming is susceptible to the errors of beamforming networks. Extensive simulations on the simplified multi-tone distributed transmit beamforming models have been conducted and offered some guidelines for system designs. Distributed beamforming for transmission of other wider band occupying signals with various data rates in frequency selective fading channels is also of our interest and will be investigated in the future.

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