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Efficacy of Hybrid Energy Beamforming with Phase Shifter Impairments and Channel Estimation Errors

Deepak Mishra, *Member, IEEE*, and Håkan Johansson, *Senior Member, IEEE*

Abstract—Hybrid energy beamforming (HEB) can reduce the hardware cost, energy consumption, and space constraints associated with massive antenna array transmitter (TX). With a single radio frequency (RF) chain having N digitally controlled phase shifter pairs, one per antenna element, theoretically achieving the same performance as a fully digital beamforming architecture with N RF chains, this letter investigates the practical efficacy of the HEB. First adopting the proposed analog phase shifter impairments model and exploiting the channel reciprocity along with the available statistical information, we present a novel approach to obtain an accurate minimum mean-square error estimate for the wireless channel between TX and energy receiver (RX). Then, tight analytical approximation for the global optimal time allocation between uplink channel estimation and downlink energy transfer operations is derived to maximize the mean net harvested energy at RX. Numerical results, validating the analysis and presenting key design insights, show that with an average improvement of 58% over the benchmark scheme, the optimized HEB can help in practically realizing the fully digital array gains.

Index Terms—Wireless power transfer, analog beamforming, MISO channel estimation, hardware impairments, optimization

I. INTRODUCTION

Massive antenna array transmitter (TX) can power the miniature wireless energy harvesting (EH) devices in internet-of-things (IoT) [1]. However, the major underlying bottlenecks are large physical size of antenna array at usable radio frequency (RF) [2], and increased signal processing cost due to the digital precoding over hundreds of antennas [3]. To address them, there has been a growing research interest [2]–[13] on investigating the hybrid beamforming architectures, where all or a part [4] of the processing is based on analog beamforming, requiring substantially fewer RF chains than the array size.

A. State-of-the-Art

In the seminal works [5], [6], it was proved that with two digitally controlled phase shifters (DCPSs) for each antenna element, the corresponding hybrid energy beamforming (HEB) with single RF chain can achieve the same performance as in a fully digital system with each antenna element having its own RF chain. This chain includes many active components like low-noise amplifier, frequency up-converter, and digital-to-analog converter [14]. Whereas, analog beamforming architectures [15] can be implemented at much lower cost and smaller form-factor using the DCPSs. However, the channel estimation (CE) for implementing HEB is more challenging because the

Fig. 1. Adopted system model for the DL hybrid RF energy beamforming from TX using UL channel estimation via pilot signal transmission from RX.

underlying effective channel is the product of random fading gain and selected analog beams [2]–[13]. To address this, an adaptive compressed-sensing based CE was studied in [7] for a massive hybrid architecture. Considering a multiuser hybrid system, a minimum mean-square error (MMSE) based CE was conducted in [8]. Joint least-squares based CE [9] and analog beam selection was proposed in [10] for the uplink (UL) multiuser hybrid beamforming system. More recently, a new CE approach for the hybrid architecture-based millimeter wave systems was proposed in [11]. As obtaining full channel state information (CSI) is difficult due to much lesser RF chains, a low-complexity hybrid precoding was implemented in [12] via beam searching in the downlink (DL). To alleviate these signaling costs, a single-stage feedback scheme for the digital precoder designing was also recently proposed in [13].

B. Motivation and Contributions

The existing works on CE for hybrid beamforming architectures [7]–[13] focused on efficient spatial multiplexing and *did not investigate the joint optimal CE and resource allocation* for maximizing HEB gains. Further, due to the usage of low-quality RF components for the ubiquitous deployment of EH devices in IoT while making massive antenna array system economically viable, the performance of these energy sustainable systems is *more prone to RF imperfections* caused by real phase shifters (PS) and losses due to practical combiners [16]–[19]. With the architecture shown in Fig. 1, this work aims at investigating the optimal average harvested energy at receiver (RX) due to HEB as compared to the fully digital ones [20] for practical massive multiple-input-single-output (MISO) energy transfer (ET) from the TX. The key contribution is four-fold:

- We introduce a *practical model* for incorporating the analog phase shifter impairments (API) in HEB architecture.
- *Novel antenna-switching based MMSE estimator* for the effective channel is proposed to study the impact of API.

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- Tight analytical approximation for the *global optimal time* allocation (TA) between the CE and ET phases is derived.
- Numerical results *validate* the proposed analysis, provide design insights, and *quantify* the achievable HEB gains.

II. SYSTEM DESCRIPTION

A. Nodes Architecture and Massive MISO Channel Model

We consider massive MISO RF-ET from a TX, consisting of single RF chain shared among all N antenna elements, to an EH RX having single antenna. As depicted in Fig. 1, this RX can be a low-power RF-EH IoT device programmed for performing an application-specific operation using its own micro-controller (μC). With $N \gg 1$, we assume flat quasi-static Rayleigh block fading [21, Ch. 2.2] where the channel impulse response for each communication link remains constant during a coherence interval of τ seconds (s), and varies independently across the different coherence blocks. The TX-to-RX channel is represented by a zero mean circularly symmetric Gaussian complex random vector, $\mathbf{h} \in \mathbb{C}^{N \times 1}$, having the variance β that incorporates both the fading gain and propagation losses [21].

B. Adopted Hybrid Energy Beamforming Protocol

For efficient DL ET, the HEB implementation at TX using one RF chain with $2N$ DCPSs [5], (i.e., one DCPS pair with a combiner per antenna element) can achieve the same array gains as a fully digital architecture with N RF chains. However, this is true only when perfect CSI is available at the TX. In general, though, as CSI is only known with certain uncertainty, the CE protocol for HEB is more challenging to implement as it involves both the CE and analog beam selection [2]–[13]. Nevertheless, for a single antenna RX with no hardware impairments in the DCPSs implementation, the CE for HEB can be effectively designed as discussed next.

Assuming channel reciprocity due to the adoption of time-division duplexing [20], the DL channel coefficients are obtained by estimating them from the UL pilot signal transmission from the RX to TX. With the HEB protocol as depicted in Fig. 1, each coherence interval of τ s is divided into two sub-phases: (a) UL CE and (b) DL ET. During the CE phase of $N\tau_c$ duration, the RX sends a pilot signal to the TX. For remaining block duration $\tau - N\tau_c$, the TX aims at focusing most of its power in the direction of RX by implementing maximum-ratio-transmission (MRT) using the obtained MMSE estimate.

Before discussing the MMSE-based novel CE approach and the corresponding TA for optimizing the DL ET to RX in Sections IV and V, we next present the proposed API model for incorporating practical limitations in implementation of HEB architecture at TX using N DCPS and combiner pairs.

III. ANALOG PHASE SHIFTER IMPAIRMENTS MODEL

Using [5, Theorem 1], a complex number $a = |a| e^{j\angle a}$ with $|a| \leq 2$, can be practically implemented using two DCPSs as:

$$a = e^{j(\cos^{-1}(\frac{|a|}{2}) + \angle a)} + e^{-j(\cos^{-1}(\frac{|a|}{2}) - \angle a)}. \quad (1)$$

This maximum amplitude constraint can be moved to the more versatile digital block [6]. Hence, in the analog block (cf.

Fig. 1), each antenna element has two DCPSs and two combiners, one as input and the other as output. For example, two practical HEB architectures are presented in [22]. The DCPSs and combiners in these architectures suffer from amplitude and phase errors due to several practical constraints like finite resolution PSs, mismatches in PS circuit elements, and channel uncertainty limiting the precision [17]–[19]. These errors cause an imbalance in the DCPS pair for each antenna element. Their sources are generally random, but some are fixed depending on the manufacturing errors and long-term aging effects. Consequently, the unpredictable and time varying API [17]–[19] can be modeled using random variables, with fixed errors deciding their means, and the randomness controlled via variances.

Following the representation in (1) and Fig. 1, we propose to model the complex baseband equivalent signal, under the API in the practical implementation of DCPS, as defined below:

$$\tilde{a} = \Theta \{a\} \triangleq g_{A_1} e^{j(\cos^{-1}(\frac{|a|}{2}) + \angle a + \phi_{A_1})} + g_{A_2} e^{-j(\cos^{-1}(\frac{|a|}{2}) - \angle a - \phi_{A_2})}, \quad (2)$$

where g_{A_1} and g_{A_2} respectively represent amplitude errors due to API at the first and second DCPS in a pair. Whereas, ϕ_{A_1} and ϕ_{A_2} respectively represent the corresponding phase errors. For the ideal case with no API, $g_{A_1} = g_{A_2} = 1$, $\phi_{A_1} = \phi_{A_2} = 0^\circ$.

We assume that these random errors in amplitude and phase for each DCPS pair and combiner are independently distributed across all TX antenna elements. With these API being uniformly distributed around fixed manufacturing error, the amplitude and phase errors are respectively modeled as:

$$g_{A_i} \triangleq 1 + \Delta_i + \Psi_i, \quad \text{and} \quad \phi_{A_i} \triangleq \Delta_i + \Psi_i, \quad \forall i = 1, 2, \quad (3)$$

where, the constant $\Delta_i > 0$ represents the errors due to the fixed sources, and Ψ_i representing the errors due to the random sources is assumed to follow the uniform distribution with the probability density function $f_{\Psi_i}(x) = \frac{1}{\delta_i}, \forall x \in [-\frac{\delta_i}{2}, \frac{\delta_i}{2}]$. Thus, δ_i controls the degree of randomness around fixed Δ_i .

IV. PROPOSED MMSE BASED CHANNEL ESTIMATION

Here, after presenting a *novel* antenna switching based analog CE to address single RF chain constraint, we present the digital channel estimator minimizing the underlying MMSE.

A. Novel Antenna-Switching Based Analog Channel Estimator

With RX transmitting continuous-time pilot signal $s(t)$ satisfying $\int_0^{N\tau_c} |s(t)|^2 dt = 1$, the received baseband signal at TX is:

$$\mathbf{y}(t) = \sqrt{E_c} \mathbf{h} s(t) + \mathbf{w}(t), \quad \forall t \in [0, N\tau_c]. \quad (4)$$

Here $E_c \triangleq p_c N\tau_c$ is the energy spent in Joule (J) at RX with p_c denoting its transmit power during CE and $\mathbf{w}(t) \in \mathbb{C}^{N \times 1}$ representing received additive white Gaussian noise (AWGN).

Letting $\mathbf{f}_A \in \mathbb{C}^{N \times 1}$ denote the analog channel estimator, the received signal available for digital processing is given by:

$$\mathbf{y}_A(t) = \mathbf{f}_A^T(t) \mathbf{y}(t), \quad \forall t \in [0, N\tau_c]. \quad (5)$$

Since there is only one RF chain to estimate N entries of the channel vector $\mathbf{h}$, we propose RX to repeatedly transmit the pilot signal over N slots of τ_c duration each. Consequently, the

Fig. 2. Graphical interpretation of the proposed analog CE under no API.

proposed analog estimator is set such that each CE phase of $N\tau_c$ duration can be sub-divided into N sub-phases. During the k th CE sub-phase interval $\tau_{ck} \triangleq ((k-1)\tau_c, k\tau_c]$, the corresponding analog estimator, as denoted by $\bar{\mathbf{f}}_{A_{id}}$, is set as:

$$[\bar{\mathbf{f}}_{A_{id}}]_i = \begin{cases} 1, & i = k, \\ 0, & i \neq k, \end{cases} \quad \forall i \in \mathcal{N} \triangleq \{1, 2, \dots, N\}. \quad (6)$$

With the above definition, the proposed analog CE process under no API, as graphically represented in Fig. 2, reduces the parallel estimation of N entries of vector $\mathbf{h}$ over a duration of $N\tau_c$ s to a sequential estimation of each entry $[\mathbf{h}]_i, \forall i \in \mathcal{N}$, each over τ_c s duration. In other words, the analog CE phase can be observed as an *antenna-switching* approach involving a sequence of N CE sub-phases, with TX only having the i th antenna active during the i th sub-phase. Thus, the MISO CE eventually reduces to a single-input-single-output CE process, but at the cost of N fold increase in the otherwise CE duration.

Despite that, since the practical HEB implementation suffers from API as discussed in Section III, the realistic entries of the analog estimator $\mathbf{f}_A(t)$ for $t \in ((i-1)\tau_c, i\tau_c], \forall i \in \mathcal{N}$, which remain the same for each τ_c duration are defined below:

$$[\bar{\mathbf{f}}_A]_i \triangleq \Theta\{[\bar{\mathbf{f}}_{A_{id}}]_i\}, \quad \forall i \in \mathcal{N}, \quad (7)$$

where $\Theta\{\cdot\}$ is as defined in (2). Substituting (6) in (7) and simplifying, the combined signal in the form of N vectors as received over N slots with interval $t \in ((i-1)\tau_c, i\tau_c], \forall i \in \mathcal{N}$, the entries of analog estimator matrix $\mathbf{F}_A \in \mathbb{C}^{N \times N}$ are:

$$[\mathbf{F}_A]_{ik} = \begin{cases} \frac{g_{A_{i1}} e^{j\phi_{A_{i1}}} (1+j\sqrt{3}) + g_{A_{i2}} e^{j\phi_{A_{i2}}} (1-j\sqrt{3})}{2}, & i = k, \\ j(g_{A_{i1}} e^{j\phi_{A_{i1}}} - g_{A_{i2}} e^{j\phi_{A_{i2}}}), & i \neq k. \end{cases} \quad (8)$$

Hence, with the above proposed analog estimator, the corresponding signal $\mathbf{y}_A(t) \in \mathbb{C}^{N \times 1}$ received as an input to digital channel estimator block, and obtained after using (8) in (5) is:

$$[\mathbf{y}_A]_i(t) = \sum_{k=1}^N [\mathbf{F}_A]_{ki} [\mathbf{y}]_k(t) = \sqrt{E_c} [\mathbf{h}_A]_i s(t) + [\mathbf{w}_A]_i(t), \quad \forall \{t \in ((i-1)\tau_c, i\tau_c]\} \wedge \{i \in \mathcal{N}\}, \quad (9)$$

where $\mathbf{h}_A \triangleq \mathbf{F}_A^T \mathbf{h}$ and $\mathbf{w}_A(t) \triangleq \mathbf{F}_A^T \mathbf{w}(t)$. For ideal (no API) scenario, $\mathbf{F}_A = \mathbf{I}_N$ with $g_{A_{ij}} = 1$ and $\phi_{A_{ij}} = 0^\circ, \forall i \in \mathcal{N}, j = 1, 2$.

B. MMSE Digital Channel Estimator

On applying MMSE minimization criterion [23] to signal $\mathbf{y}_A(t)$, the resulting digital channel estimator is given by:

$$\mathbf{f}_D(t) \triangleq \frac{\text{cov}\{\mathbf{h}_A, \mathbf{y}_A\}}{\text{cov}\{\mathbf{y}_A, \mathbf{y}_A\}} \mathbf{y}_A(t) \stackrel{(r1)}{\approx} \frac{\beta \sqrt{E_c} \mathbf{y}_A(t)}{\beta E_c + \sigma_w^2}, \quad (10)$$

where (r1) is obtained on applying no API assumption to covariance definitions [23], and σ_w^2 denotes AWGN variance. Using (10), MMSE estimate $\hat{\mathbf{h}}_A$ for effective channel $\mathbf{h}_A$ is:

$$\hat{\mathbf{h}}_A = \int_0^{N\tau_c} \frac{s^*(t)}{\sqrt{E_c}} \mathbf{f}_D(t) dt \approx \hat{\mathbf{h}}, \quad (11)$$

where $\hat{\mathbf{h}}$ is the MMSE estimate for $\mathbf{h}_A$ under no API assumption. However, for the practical HEB with API, $\hat{\mathbf{h}}$ is an approximation for $\hat{\mathbf{h}}_A$, and we have used it in next section for obtaining the optimal CE time that maximizes the ET gains.

V. JOINT OPTIMAL PRECODING AND TIME ALLOCATION

Using $\hat{\mathbf{h}}_A$, the optimal precoder for HEB should be designed such that it maximizes the received power at RX by focusing most of the TX's radiated power in the direction of RX. Hence, to maximize ET gains, the *MRT-based precoder* design needs to be selected at TX by setting the digital precoder as $\mathbf{z}_D \triangleq \sqrt{p_d}$, with the analog precoder being $\mathbf{z}_A \triangleq \frac{\hat{\mathbf{h}}_A}{\|\hat{\mathbf{h}}_A\|} \in \mathbb{C}^{1 \times N}$. Here, p_d is the TX power of TX during DL RF-ET and $\hat{\mathbf{h}}_A$ is the MMSE for the channel $\mathbf{h}$. However, under API, the analog precoder $\mathbf{z}_A$ gets practically altered to $\bar{\mathbf{z}}_A$, as defined below:

$$[\bar{\mathbf{z}}_A]_i \triangleq \Theta\{[\mathbf{z}_A]_i\}, \quad \forall i \in \mathcal{N}. \quad (12)$$

A. Average Net Harvested Energy at RX During ET Phase

Using (12), together with the API model in (2), the average power $\mu_{pr} = \mathbb{E}\{|\mathbf{z}_D \bar{\mathbf{z}}_A \mathbf{h}|^2\}$ at RX can be approximated as:

$$\mu_{pr} \stackrel{(r2)}{\approx} p_d \mathbb{E}_{\mathbf{h}} \left\{ \frac{\hat{\mathbf{h}}^H \mathbb{E}\{\mathbf{h} \mathbf{h}^H | \hat{\mathbf{h}}\} \hat{\mathbf{h}}}{\|\hat{\mathbf{h}}\|^2} \right\} \stackrel{(r3)}{=} p_d \beta \left[\frac{N \beta E_c + \sigma_w^2}{\beta E_c + \sigma_w^2} \right]. \quad (13)$$

where (r2) is obtained on employing the no API assumption (i.e., $\mathbf{F}_A = \mathbf{I}_N$) and (r3) on using $\text{cov}\{\mathbf{h} | \hat{\mathbf{h}}\} = \frac{\beta \sigma_w^2}{\beta E_c + \sigma_w^2} \mathbf{I}_N$, $\mathbb{E}\{\mathbf{h} | \hat{\mathbf{h}}\} = \frac{\beta E_c \hat{\mathbf{h}}}{\beta E_c + \sigma_w^2}$, with $\mathbb{E}_{\mathbf{h}}\{\|\hat{\mathbf{h}}\|^2\} = N \left(\beta + \frac{\sigma_w^2}{E_c} \right)$.

Since the RX spends E_c amount of energy during the CE phase, its net harvested energy E_n during a coherence block with η denoting rectification efficiency of the EH unit [24] is:

$$E_n \triangleq \eta (\tau - N\tau_c) |\mathbf{z}_D \bar{\mathbf{z}}_A \mathbf{h}|^2 - E_c. \quad (14)$$

Using approximation (13) to average power received during ET in (14), average net harvested energy is approximated as:

$$\mathbb{E}\{E_n\} \approx \mu_{E_n} \triangleq \eta (\tau - N\tau_c) p_d \beta \left[\frac{N^2 \beta p_c \tau_c + \sigma_w^2}{N \beta p_c \tau_c + \sigma_w^2} \right] - N p_c \tau_c. \quad (15)$$

B. Analytical Approximation for Optimal CE Time Allocation

With approximation for net harvested energy given in (15) as the objective to be maximized by optimally allocating the available time between the CE and ET phases is given below.

Lemma 1: The global optimal TA for CE phase is given by:

$$\hat{\tau}_c^* \triangleq \max \left\{ 0, \frac{\sigma_w^2}{N \beta p_c} \left[\sqrt{\left(1 + \frac{\beta p_c \tau}{\sigma_w^2} \right) \frac{\eta p_d \beta (N-1)}{p_c + \eta p_d \beta N}} - 1 \right] \right\}. \quad (16)$$

Proof: From $\frac{\partial^2 \mu_{E_n}}{\partial \tau_c^2} = -\frac{2\beta^2 \eta (N-1) \sigma_w^2 p_d p_c (\sigma_w^2 + \beta p_c \tau)}{(\sigma_w^2 + N \beta p_c \tau_c)^3 (N)^{-2}} \leq 0$, $\forall N \geq 1$, we notice the *concavity* [25] of μ_{E_n} in τ_c . Hence, the

Fig. 3. Validation of the proposed CE analysis and HEB optimization.

global optimal time $\hat{\tau}_c^*$ for CE maximizing μ_{E_n} (or providing approximate for maximizer of $\mathbb{E}\{E_n\}$), as obtained by solving $\frac{\partial \mu_{E_n}}{\partial \tau_c} = 0$ in $\tau_c \in [0, \frac{\tau}{N}]$, yields the desired result in (16). ■ The accuracy of this proposed analytical solution $\hat{\tau}_c^*$ has been numerically validated against the one maximizing $\mathbb{E}\{E_n\}$.

VI. NUMERICAL PERFORMANCE EVALUATION

Here, we numerically evaluate the optimized performance of the proposed HEB protocol under API and CE errors. The default parameter values are set as: $N = 20$, $\tau = 10^{-2}\text{s}$, $p_d = 4\text{W}$, $p_c = 10^{-3}\text{W}$, $\sigma_w^2 = 10^{-20}\text{J}$, $\eta = 0.7$, and $\beta = G\varpi d_i^{-\varrho}$, where $\varpi = (\lambda/(2\pi))^2$ being the average attenuation at unit reference distance with $\lambda = 3 \times 10^8(2f_c)^{-1}$, $f_c = 915\text{MHz}$ [24], $d = 5\text{m}$, $\varrho = 2.5$ as path loss exponent, and $G = 1$ implying omni-directional antenna elements. Regarding API parameters for each DCPS pair and combiner, we consider $\delta_i = -\Delta_i = \Delta$, $\forall i = 1, 2$. Further, generally, Δ takes low value, i.e., $\Delta < 0.1$, because for the practical PSs design [18], [19], $\phi_{A_i} \ll 10^\circ$. Thus, we use $\Delta = 0.06$ which implies $\phi_{A_i} \leq 5^\circ$. Moreover, to corroborate the utility of optimal TA (OTA) $\hat{\tau}_c^*$, we use fix TA as $\tau_{c_0} \triangleq \frac{\tau}{100N}$ to ensure $\mathbb{E}\{E_n\} \geq 0$. Lastly, all the simulation results plotted here have been obtained after taking average over the 10^5 independent channel realizations.

First via Fig. 3(a) we validate the quality of the proposed MMSE estimate $\hat{\mathbf{h}}$ defined in (11). With average net harvested energy $\mathbb{E}\{E_n\}$ as the performance validation metric for estimating the goodness of $\hat{\mathbf{h}}$, we have also plotted the perfect CSI (ideal) and isotropic [26, Ch. 2.2] transmission cases to respectively give the upper and lower bounds on $\mathbb{E}\{E_n\}$ for the fully digital architecture. As observed from Fig. 3(a), the quality of the proposed MMSE improves with increasing signal-to-noise ratio (SNR) $\gamma_E \triangleq \frac{E_c\beta}{\sigma_w^2}$ because the underlying CE error reduces, and for higher SNRs satisfying $\gamma_E > 17.5\text{dB}$, the corresponding $\mathbb{E}\{E_n\}$ approaches the ideal digital performance, with a negligible gap owing to non-zero CE time $\tau_c = \tau_{c_0}$. Further, for SNRs $\gamma_E < -17.5\text{dB}$, the isotropic transmission is better than HEB because the underlying harvested energy during ET is not sufficiently large in comparison to consumption E_c at RX for implementing CE.

Now, we validate the analytical claims in Lemma 1 regarding the concavity of $\mathbb{E}\{E_n\}$ (or μ_{E_n}) in τ_c along with the accuracy of the approximation $\hat{\tau}_c^*$ for the global OTA. From Fig. 3(b), where $\mathbb{E}\{E_n\}$ (cf. (15)) is plotted against increasing τ_c for different TX-to-RX distance d values, it can be observed that $\mathbb{E}\{E_n\}$ is unimodal in τ_c . Here, $\mathbb{E}\{E_n\} = 0$ for $\tau = N\tau_c$,

Fig. 4. Performance comparison and impact of API on HEB gains.

and the value at $\tau_c = 0$ represents isotropic performance. Lastly, it is worth noting that the proposed approximation $\hat{\tau}_c^*$ provides a very tight match to the numerical global optimal τ_c^* , as obtained via an exhaustive search, especially for larger d values. Furthermore, as for lower SNRs, more time needs to be allocated for accurate CE, $\hat{\tau}_c^*$ is higher for larger d values.

Next, we quantify the achievable gains using the OTA $\hat{\tau}_c^*$ by plotting $\mathbb{E}\{E_n\}$ due to the proposed HEB with increasing γ_E in Fig. 4(a) for both the fix TA τ_{c_0} and OTA $\hat{\tau}_c^*$. The corresponding values of $\mathbb{E}\{E_n\}$ for the fully digital architecture with OTA and benchmark HEB scheme as proposed in [8] are also plotted for comparison. The benchmark scheme is based on designing the analog channel estimator by selecting first α left-singular vectors of the MMSE estimator for fully digital architecture. As in general, the optimal $\alpha \leq \frac{N}{2}$ [8, Section VI-A.1] we have set $\alpha = 10$. With a less than 1.6% of negligible degradation in net harvested energy as compared to the fully digital beamforming with N RF chains, the proposed single RF chain based HEB protocol with OTA for MMSE-based CE yields a significant average performance gain of around 58% over the existing benchmark. Further, the proposed OTA $\hat{\tau}_c^*$ provides an enhancement of about 37% over an arbitrary fix TA τ_{c_0} , which performs even poorer than benchmark scheme for $\gamma_E \leq -17.5\text{dB}$ due to the relatively longer CE for $\tau_c = \tau_{c_0}$.

Lastly, we show the impact of API on the HEB performance both with OTA and fix TA for $\gamma_E = \{-10, 20\}\text{dB}$ in Fig. 4(b). As the API parameter Δ increases from 0 (no impairment) to 0.1 (maximum degradation [18] having $\phi_{A_i} \approx 10^\circ$), the underlying net mean harvested energy for $\tau_c = \hat{\tau}_c^*$ and $\tau_c = \tau_{c_0}$ respectively decreases by 10% and 5% for $\gamma_E = -10\text{dB}$. Whereas, for higher SNRs like $\gamma_E = 20\text{dB}$, where OTA and fix TA have similar HEB performance, this degradation is $\approx 7\%$.

VII. CONCLUDING REMARKS

This letter investigated the practical efficacy of using a single RF chain at a massive antenna array TX in wirelessly delivering energy to an EH RX. Firstly, a practical API model was introduced to characterize the hardware impairments in DCPS implementation. Thereafter, the MMSE estimate of the underlying effective channel was obtained using the novel antenna-switching proposal and analytical approximation to OTA for CE was derived. Our extensive simulations quantifying the array gains over benchmarks corroborated the practical utility of the proposed cost and space-effective optimized green HEB protocol. In the future, we would like to characterize practical utility of HEB over frequency-selective channels.

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