Received October 31, 2018, accepted November 23, 2018, date of publication December 17, 2018, date of current version January 29, 2019.

Digital Object Identifier 10.1109/ACCESS.2018.2886297

Receiver Design to Employ Simultaneous Wireless Information and Power Transmission with Joint CFO and Channel Estimation

AKASHKUMAR RAJARAM[1,2,3], (Student Member, IEEE), RUI DINIS[2,3], (Senior Member, IEEE), DUSHANTHA NALIN K. JAYAKODY[1], (Senior Member, IEEE), AND NEERAJ KUMAR[4], (Senior Member, IEEE)

[1]School of Computer Science and Robotics, National Research Tomsk Polytechnic University, 634050 Tomsk, Russia
[2]Faculty of Science and Technology, New University of Lisbon, 1049-001 Lisbon, Portugal
[3]Instituto de Telecomunicações, 1049-001 Lisbon, Portugal
[4]Computer Science and Engineering Department, Thapar Institute of Engineering and Technology, Patiala 147003, India

Corresponding author: Dushantha Nalin K. Jayakody (nalin@tpu.ru)

This work was supported in part by the Framework of the Competitiveness Enhancement Program of the National Research Tomsk Polytechnic University, Russia, and in part by the Fundação para a Ciência e Tecnologia and the Instituto de Telecomunicações under Project UID/EEA/50008/2013.

ABSTRACT Radio-frequency energy harvesting (EH) is one of the enabling technologies for the next-generation wireless communication systems. EH techniques are specifically used to improve the energy efficiency of the system. Recently, the simultaneous wireless information and power transmission (SWIPT) protocol is adapted for EH. In this paper, we design a new receiver for joint carrier frequency offset (CFO) and channel estimation on single-carrier modulations with frequency-domain equalization along with SWIPT implementation for EH by using the pilot signal. The pilot signal is a highly energized signal, which is superimposed with the information signal. The superimposed signal is used not only to transmit power for EH purposes but also to estimate the CFO and channel conditions. The receiver is designed to accommodate the strong interference levels in the channel estimation and data detection. The proposed scheme offers a flexible design method and efficient resource utilization. We validate our analytical results using simulations.

INDEX TERMS Channel estimation, CFO estimation, energy harvesting, SWIPT, SC-FDE.

I. INTRODUCTION

The concept of energy efficiency is a concrete ground for energy aware 5^{th} generation communications (5G) and "Smart Cities". The concept of wireless Radio Frequency (RF) energy harvesting (EH) is gaining importance due to the limitation of conventional EH techniques. RF-EH has recently regarded as a promising avenue for energy-constrained wireless networks [1], [2]. There are numerous research articles which are focused on improving RF-EH by efficient usage of available resources [1], [3]. One of the basic RF-EH technique is wireless power transfer (WPT) and in this technique, the RF-EH is supported by a dedicated energy source and/or opportunistically EH from ambient RF signals. WPT is further evolved to an another technique called the simultaneous wireless information and power transfer (SWIPT) [2]. SWIPT is one of the promising RF-EH technique, because in this technique both information and energy can be transmitted simultaneously, thus saving spectral resources. This results in considerable gains in terms of spectral efficiency and energy consumption.

In this paper, we use single-carrier frequency-division multiple access (SC-FDMA), SC-FDMA has several advantages over orthogonal frequency-division multiple access (OFDMA) as mentioned in [4]. But it also has a main disadvantage as compared to OFDMA i.e. occurrence of carrier frequency offset (CFO) during signal transmission. SC-FDMA is sensitive to CFO as compared to OFDMA [4]. CFO occurs mainly due to the frequency mismatch between the oscillators at the transmitter and at the receiver [5], [6], and due to Doppler shift, and the compensation techniques are studied in [7]. Thus, it is important to study CFO effects and the compensation technique for the SC-FDMA signal.

Frequency errors due to CFO directly affect the performance of channel estimation and signal detection [8], [9]. There are many CFO estimation techniques proposed for OFDM schemes in [10]–[12]. Maximum likelihood frequency offset estimation technique was proposed in [10] and this method is suitable for small CFO, because it compares two consecutive and identical symbols with the symbol duration T and the frequency acquisition range is $\pm 1/(2T)$. To improve the acquisition range for the CFO, two separate pilot signals were employed in [11]. Based on [11], an algorithm called best linear unbiased estimator (BLUE) is proposed in [12] to improve the acquisition range.

To improve information decoding accuracy at receiver, an efficient iterative frequency domain equalization (FDE) for SC-FDE is introduced [13]. It is called as iterative block decision feedback equalization (IB-DFE) and this technique performs better than non-iterative methods [13]. Recently, an iterative linear minimum mean-square-error is proposed to estimate the individual channels used in multiple input and multiple antenna system [14]. Thus, the iterative block based receiver structure is a predominantly researched area, especially, the IB-DFE receiver. By using IB-DFE receiver, the channel estimation is performed with the help of pilot signal as in [15]. Since the optimum FDE coefficients are a function of the channel frequency response, channel estimates are required at the receiver.

In this paper, we use Moose technique, assuming that the frequency acquisition range is within $\pm 1/(2T)$ and CFO is small. Also we the idea of using SWIPT in the point to point communication model of 5G networks and it is applicable to both the uplink and downlink. The 5G network uses both the single carrier (SC) modulation and the multi carrier (MC) modulation. The SC for uplink and MC modulation for downlink as it proved to be the best choice [16]. The SC-FDMA with Frequency-Domain Equalization (FDE) is found to be suitable for the transmission of high data rate signal over severely time dispersive channels.

The channel estimates are obtained by using pilot symbols, which are either in time or frequency domain [17]. Most commonly, frequency domain is used for OFDM modulations while both time and frequency domain is used for SC modulations. In block transmission techniques, the channel impulse response maybe very long and the record over-heads for channel estimation is possibly very high. As a solution, the pilot symbols are superimposed with the information symbols,instead of multiplexing the pilot symbols with the information symbols. This increases the density of pilots with respect to information symbols without comprising the spectral efficiency. This helps in improving the amount of EH. The disadvantages of using superimposed pilot signal, is the interference of pilot signal with the information signal with increase in transmit power of pilot signal. Then, the interference of pilot signal with information signal can be reduced by averaging the channel estimate of the respective frequency. The proposed technique helps in estimating CFO as shown in [10] by using the highly energized identical pilot signal sequence on each block. Thereby with help of SWIPT, pilot signal will be more robust for signal interference and noise.

The SWIPT improves the performance of the iterative receiver by allowing IB-DFE in exploiting the excessive power, that was originally intended for EH, to estimate channel. This, recursively, can be used in the iterative receiver to decode. This reduces the nonlinear distortion effects and minimize the estimation overheads [18]. The effective use of SWIPT with IB-DFE not only increases the spectral efficiency, but also reduces the signal interference and helps in signal detection.

The contribution of this paper is three fold,

- we design a receiver for SWIPT with joint CFO and channel estimation. The pilot symbols, which are superimposed, with the information symbols are used for EH, CFO and channel estimation. Here, the overall transmit power of the pilot signal is higher than that of other competitive models such as techniques which employs multiplexing pilot symbols with the information symbols.
- we improve the channel estimation accuracy with the help of IB-DFE and an algorithm is introduced with IB-DFE to increase the accuracy of the channel estimation with the estimates feedback in IB-DFE.
- we find the minimum power required for the pilot signal to estimate the channel condition and CFO with achievable accuracy for the given power of the information signal. Also, we find the optimum power allocation ratio between the information and the pilot signals by using simulation results.

The structure of the paper as follows. The system model is explained in Sec. II. The proposed channel estimation technique and signal detection is explained in Sec. III. The performance results are analyzed in in sec. IV and the conclusion is presented in sec. V.

II. SYSTEM MODEL

The system uses quadrature phase shift keying (QPSK) modulation with SC-FDMA over Rayleigh fast fading channel. The receiver estimates the channel and information using iterative receiver. The system harvests energy from the received signal as shown in Fig. 1. It is assumed that the system experience additive white Gaussian noise (AWGN) with variance $N_0/2$ is modeled as zero mean complex Gaussian random variable and undergoes phase rotation. The system adapts power splitting protocol based SWIPT (PS-SWIPT). The receiver fitted with a special circuit at the antenna to split the total power of the signal for information decoding and energy harvesting [1]. The frame structure of the signal is presented in Fig. 2, where time duration per symbol, block duration and total time for all the blocks are denoted as T, T_l and T_lL, respectively. Frame structure is similar to the frame structure used in [19], where the pilot and the information signals are superimposed together as a single signal. The frame structure of the signal has L number of signal blocks

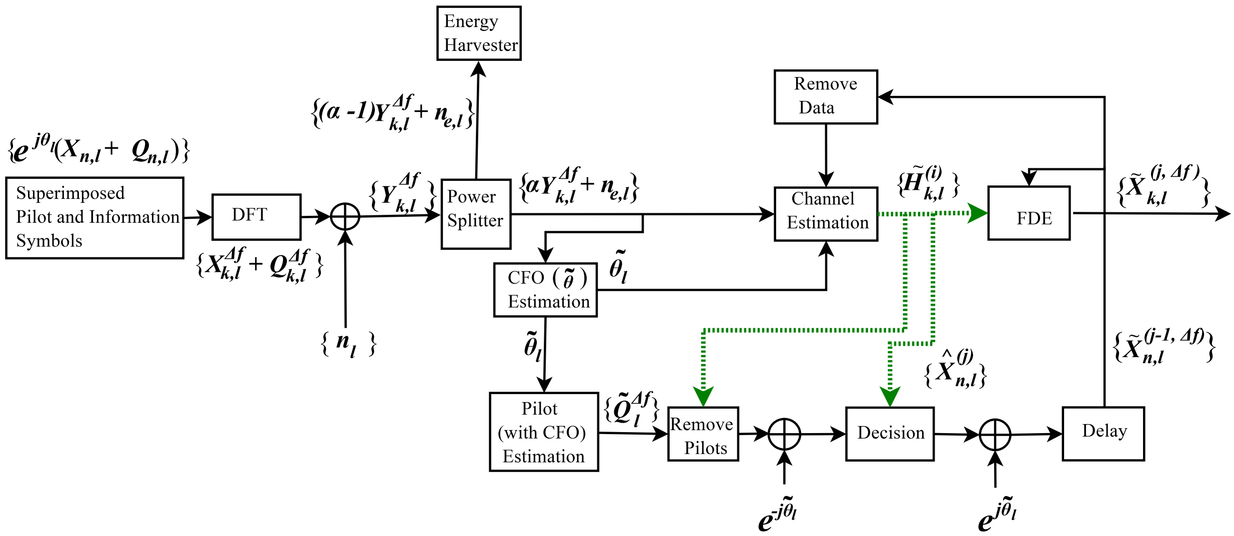

FIGURE 1. Receiver Design for SWIPT with Joint CFO and Channel Estimation.

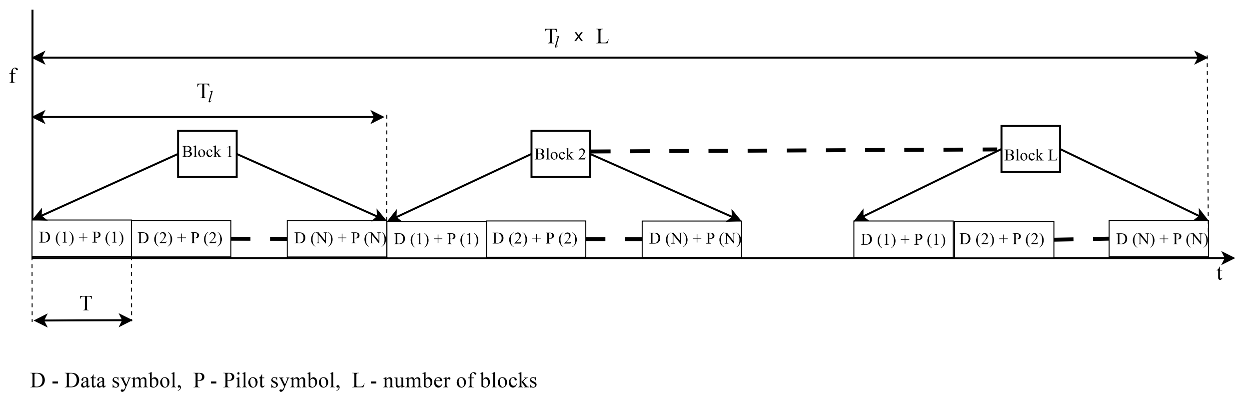

FIGURE 2. Frame structure of the proposed system, where T and T_l are time duration of a symbol and a block, respectively.

with each signal blocks has N number of symbols. The signal blocks are denoted as l, where $l \in \{0, 1, \ldots, L\}$ and the symbol is denoted as n, where $n \in \{0, 1, \ldots, N-1\}$. The received signal $'Y_l'$ can be expressed as

$$Y_l = H_l(\sqrt{P_x}X_l + \sqrt{P_q}Q_l) + W_l, \quad l \in \{0, 1, \ldots, L-1\}, \tag{1}$$

where H_l is Rayleigh fading channel, X_l and Q_l are the information signal and pilot signal with the transmit power $\sqrt{P_x}$ and $\sqrt{P_q}$, respectively. W_l is the corresponding AWGN noise.

The power allocation ratio for the information decoding (ID) and EH by using PS-SWIPT circuit are α and $(\alpha - 1)$, respectively, where $0 < \alpha < 1$. The received signal obtained from the PS-SWIPT circuit for ID and EH are denoted as $Y_{l,i}$ and $Y_{l,e}$, respectively. Thus we have

$$\begin{aligned} Y_{l,i} &= \alpha\big(H_l(\sqrt{P_x}X_l + \sqrt{P_q}Q_l) + W_l\big) + W_e, \\ Y_{l,e} &= (\alpha - 1)\big(H_l(\sqrt{P_x}X_l + \sqrt{P_q}Q_l) + W_l\big) + W_e, \end{aligned} \tag{2}$$

where W_e is the new AWGN noise occurred due to the signal splitting by power splitter circuit at the receiver [1].

The energy harvested at the receiver from the signal $Y_{l,e}$ is denoted as E_y and is written as

$$E_y = \eta_{eh}(\alpha - 1)(P_x + P_q)h_l T_l, \tag{3}$$

where η_{eh} is the EH efficiency of the rectenna and h_l is $|H_l|^2$. $|H_l|^2$ is the channel power gain of H_l with $|H_l|^2 \sim e^{\sigma_{SD}^2}$ and $\sigma_{SD}^2 = d_{SD}{}^{-\chi}\zeta_d$, where $e^{\sigma_{SD}^2}$ is exponential distribution mean of H_l. The distance between S and D is denoted as d_{SD}. The path loss factor and the power attenuation of the signal is denoted as χ and ζ_d, respectively.

The signal transmission for l^{th} block in time domain of SC-FDE modulation with appropriate phase rotation is written as

$$x_l^{Tx}(t) = \sum_{n=-Ns}^{N-1} x_{n,l}^{Tx} e^{j\theta_n} r(t - nT), \tag{4}$$

where $r(t)$, $-Ns$ are the pulse shaping filter and the cyclic prefix, respectively and $x_l^{Tx}(t)$ is the samples with time domain symbols n, $n \in \{0, 1, \ldots, N-1\}$ transmitted over a time-dispersive channel. T denotes the symbol duration. The phase rotation of each symbol present in the l is denoted as θ_n and $\theta_n = 2\pi \Delta f T_l \frac{n}{N}$, where the CFO is denoted as Δf and it is assumed that Δf is constant for all the blocks. The information signal is superimposed with the pilot signal, thus we have

$$x_{n,l}^{Tx} = e^{j\theta_n}\{\sqrt{P_x}x_{n,l} + \sqrt{P_q}q_{n,l}\}, \tag{5}$$

where n is the number of blocks symbols present in each time blocks l, with $n = 0, 1, \ldots, N-1$ and $l = 0, 1, \ldots, L-1$. Since the symbols are superimposed, θ_n affects both $x_{n,l}$ and $q_{n,l}$. Then, the superimposed time domain signal is converted to frequency domain of SC-FDE as

$$X_{k,l}^{Tx} = X_{k,l}^{(\Delta f)} + Q_{k,l}^{(\Delta f)}, \tag{6}$$

where k is the frequency of block l, $k = 0, 1, \ldots, K-1$ and $l = 0, 1, \ldots, L-1$. $X_{k,l}^{(\Delta f)} = DFT\{x_{n,l}e^{j\theta_n}; n = 0, 1, \ldots, N-1\}$ and $Q_{k,l}^{(\Delta f)} = DFT\{q_{n,l}e^{j\theta_n}; n = 0, 1, \ldots, N-1\}$ are the information and pilot signals converted from time to frequency domain, respectively.

By using (6) in (2) gives

$$Y_{k,l,i}^{(\Delta f)} = \alpha\big(H_{k,l}(\sqrt{P_x}X_{k,l}^{(\Delta f)} + \sqrt{P_q}Q_{k,l}^{(\Delta f)}) + W_l\big) + W_{e,l}. \tag{7}$$

III. JOINT CFO AND CHANNEL ESTIMATION WITH IB-DFE

In this section, the joint CFO and channel estimation along with signal detection by using iterative receiver based on IB-DFE concept is presented as shown in Fig. 1. CFO and channel estimation comprise three steps as follows:

A Compute the average CFO estimate by using $Y_{k,l,i}^{(\Delta f)}$ and CFO estimate is denoted as $\Delta\tilde{f}$. Estimate pilot signal with phase rotation by using $\Delta\tilde{f}$. The estimate of pilot signal with phase rotation is denoted as $\tilde{Q}_{k,l}^{(\Delta f)}$ and it should be used for channel estimation due to phase rotation on $Q_{k,l}$.

B To compute the average channel estimate $\tilde{H}_{k,l}^{av}$ over a set of blocks without using iterative receiver and analyze conditions involved to estimate the channel condition.

C To compute the information estimate $\tilde{X}_{k,l}^{(j,\Delta f)}$ and new channel estimates $\tilde{H}_{k,l}^{(j)}$ by using iterative receiver. In each iteration the new information estimate and channel estimate are found by using their previous value and received signals. The final information estimate found by repeating the same iterative blocks for an optimum number of times is denoted as $\tilde{X}_{k,l,F}^{(j,\Delta f)}$.

The above three steps are explained in the following subsections.

A. CARRIER FREQUENCY OFFSET ESTIMATION

CFO is estimated by calculating phase rotation over a set of received signal blocks with help of pilot signal. CFO is estimated by using Moose technique [10], [21] on the received signal. The CFO estimate of each block is written as

$$\Delta\tilde{f}_l = \frac{1}{2\pi T_l}\arg\left\{\sum_{n=0}^{N-1} y_{n,l,i}^{(\Delta f)} y_{n,(l+1),i}^{(\Delta f)\,*}\right\}, \tag{8}$$

where $*$ denotes the conjugate of $y_{n,(l+1),i}$. $Y_{k,l,i}^{(\Delta f)} = \text{DFT}\{e^{j\theta_n}\tilde{y}_{n,l,i}; n = 0, 1, \ldots, N-1\}$ and $Y_{k,(l+1),i}^{(\Delta f)} = \text{DFT}\{e^{j\theta_n}\tilde{y}_{n,(l+1),i}; n = 0, 1, \ldots, N-1\}$. The mean CFO estimate of all the blocks is written as

$$\Delta\tilde{f} = \frac{1}{L-1}\left\{\sum_{l=0}^{L-2}\Delta\tilde{f}_l\right\}, \tag{9}$$

where the phase rotation estimate is written as $e^{j\tilde{\theta}_n} = 2\pi\Delta\tilde{f}T_l\frac{n}{N}$

The expected value of $H_{k,l}$, W_l, W_e, $X_{k,l}^{(\Delta f)}$ and $Q_{k,l}^{(\Delta f)}$, respectively is written as

$$\begin{aligned}
&\mathbb{E}[H_{k,l}] = 2\sigma_{h,k,l}^2 = \mathbb{E}[W_l] = 2\sigma_{w,l}^2,\ \mathbb{E}[W_{e,l}] = 2\sigma_{w,e}^2,\\
&\mathbb{E}[X_{k,l}] = \mathbb{E}[X_{k,l}^{(\Delta f)}] = N\mathbb{E}[x_{n,l}{}^2] = 2\sigma_{x,n,l}^2,\\
&\mathbb{E}[Q_{k,l}] = \mathbb{E}[Q_{k,l}^{(\Delta f)}] = N\mathbb{E}[q_{n,l}{}^2] = 2\sigma_{q,n,l}^2,
\end{aligned} \tag{10}$$

where $\mathbb{E}[X_{k,l}] = \mathbb{E}[X_{k,l}^{(\Delta f)}]$ and $\mathbb{E}[Q_{k,l}] = \mathbb{E}[Q_{k,l}^{(\Delta f)}]$ because there is no change in variance with or without phase rotation. The expected values in (10) are used to analyze (8). To find CFO estimation from (8) and (9), the received signal should satisfy three conditions, they are:

1 At the transmitter, the pilot symbols present in each block should be same, i.e. $Q_{k,l} = Q_{k,0} = Q_{k,1}\ldots = Q_{k,L-1}$. With this condition, ideally by considering only the pilot signal (without noise and information signal), the angle between subsequent pilot blocks should give the perfect phase rotation estimate between the subsequent pilot blocks.

2 Since the CFO is constant for each symbol in the block, the phase rotation is linear and the angle between summation of all the symbols in l^{th} block and $(l+1)^{th}$ block, respectively, should give the mean phase rotation of l^{th} block.

3 At receiver, converting the received signal of l blocks from frequency to time domain, we get $e^{j\theta_n}$ of l blocks. It is written as: $y_{k,l}^{(\Delta f)} = \text{DFT}\{e^{j\theta_n}\tilde{y}_{n,l}; n = 0, 1, \ldots, N-1\}$. If $\sigma_{q,n,l}^2 > \{\sigma_{x,n,l}^2 + \sigma_{w,l}^2 + \sigma_{w,e}^2\}$ then, $e^{j\theta_n}$ of $\tilde{Y}_{n,l}$ is approximately equal to $e^{j\theta_n}$ of $q_{n,l}$. This means, if the signal strength of pilot signal is greater than information signal combined with noise in all the blocks, then $\Delta\tilde{f}$ is estimated using (8) and (9).

Due to phase rotation in the received signal, the pilot signal transmitted at the source is not used for channel estimation. Thereby, $\tilde{\theta}_n$ is used to find the pilot signal with phase rotation and it is written as

$$\tilde{Q}_{k,l}^{(\Delta f)} = \text{DFT}\{e^{j\tilde{\theta}_n}Q_{n,l}\}\ , \tag{11}$$

where $n = 0, 1, \ldots, N-1$ and $l = 0, 1, \ldots, L-1$. The estimated value of $\tilde{Q}_{k,l}^{(\Delta f)}$ is written as $\mathbb{E}[\tilde{Q}_{k,l}^{(\Delta f)}] = N\mathbb{E}[q_{n,l}{}^2] = 2\sigma_{q,n,l}^2$

B. CHANNEL ESTIMATION

The channel is estimated using the received signal and the pilot signal estimate, is written as

$$\begin{aligned}\tilde{H}_{k,l} &= \frac{Y^{(\Delta f)}_{k,l,i}}{P_q\tilde{Q}^{(\Delta f)}_{k,l}}\\ &= \alpha\left(\frac{H_{k,l}\sqrt{P_x}X^{(\Delta f)}_{k,l}+W_l}{\sqrt{P_q}\tilde{Q}^{(\Delta f)}_{k,l}}\right)+\alpha H_{k,l}+\frac{W_{e,l}}{P_q\tilde{Q}^{(\Delta f)}_{k,l}}\\ &= \alpha H_{k,l}+\varepsilon_{H\,k,l}.\end{aligned}\tag{12}$$

The channel estimate $\tilde{H}_{k,l}$ can be improved by converting $\{\tilde{H}_{k,l}; k = 0, 1, \ldots, K-1\}$ = DFT $\{\tilde{h}_{n,l}g_n; n = 0, 1, \ldots, N-1\}$ if n_{th} time domain is inside the cycle prefix, then g_n is either 1 or 0. After forcing zeros to the time domain samples which are not inside the cycle prefix, again the time domain channel estimate is converted back to frequency domain $\{\tilde{h}_{n,l}; n = 0, 1, \ldots, N-1\}$ = IDFT $\{\tilde{H}_{k,l}g_n; k = 0, 1, \ldots, K-1\}$. The channel estimate error is written as

$$\varepsilon_{H\,k,l} = \alpha\left(\frac{H_{k,l}\sqrt{P_x}X^{(\Delta f)}_{k,l}+W_l}{\sqrt{P_q}\tilde{Q}^{(\Delta f)}_{k,l}}\right)+\frac{W_{e,l}}{P_q\tilde{Q}^{(\Delta f)}_{k,l}},\tag{13}$$

where $\varepsilon_{H\,k,l}$ denotes the channel estimation error, where $\varepsilon_{H\,k,l}$ depends on the P_x, noise, P_q and pilot signal.

The frame structure is similar to [19], where each frame has N sub carriers per block. The data $X^{(\Delta f)}_{k,l}$ and the pilot signal $\tilde{Q}^{(\Delta f)}_{k,l}$ are in frequency domain and they have equal number of sub carriers. In order to reduce the envelope fluctuations of the transmitted signal, the pilot signal in time domain should be constant. Then $\tilde{Q}^{(\Delta f)}_{k,l}$ and $\tilde{Q}_{n,l}$ are constants by using Chu sequence as in [15]. By using (10) in (13) gives

$$\mathbb{E}[\varepsilon_{H\,k,l}] = \alpha\left(\frac{\sigma^2_{h,k,l}\sigma^2_{x,n,l}+\sigma^2_{w,l}}{\sigma^2_{\tilde{Q},n,l}}\right)+\frac{\sigma^2_{w,e}}{\sigma^2_{\tilde{Q},n,l}}.\tag{14}$$

The channel fades with the change in frequency, but remains constant for the respective frequency of a set of transmitted signal. Then based on this assumption, averaging the channel estimation reduces the error and then (14) can be is written as

$$\varepsilon_{H^{av}_{k,l}} = \frac{1}{l}\sum_{l=0}^{L-1}\left(\mathbb{E}[\varepsilon_{H\,k,l}]\right),\quad l=0,1,\ldots,L-1,\tag{15}$$

$$\tilde{H}^{av}_{k,l} = \alpha(H_{k,l}+\varepsilon_{H^{av}_{k,l}}),\tag{16}$$

where the average of channel estimation error and channel estimation over l blocks are denoted as $\varepsilon_{H^{av}_{k,l}}$ and $\tilde{H}^{av}_{k,l}$, respectively.

Thus, the information estimated by using the received signal, pilot signal and the channel estimate $\tilde{H}^{av}_{k,l}$ is written as

$$\begin{aligned}\tilde{X}^{(z,\Delta f)}_{k,l} &= \frac{Y^{(\Delta f)}_{k,l,i}}{\tilde{H}^{av}_{k,l}}-\alpha\sqrt{P_q}\tilde{Q}^{(\Delta f)}_{k,l}\\ &= \frac{\alpha\left(H_{k,l}(\sqrt{P_x}X^{(\Delta f)}_{k,l}+\sqrt{P_q}\tilde{Q}^{(\Delta f)}_{k,l})+W_l\right)+W_{e,l}}{\alpha(H_{k,l}+\varepsilon_{H^{av}_{k,l}})}\\ &\quad -\alpha\sqrt{P_q}\tilde{Q}^{(\Delta f)}_{k,l},\end{aligned}\tag{17}$$

where $\tilde{X}^{(z,\Delta f)}_{k,l}$ is the information estimate computed by using zero forcing decoder. The expected value of information estimate error is calculated by using mean square error (MSE) and it is written as

$$\mathbb{E}[\varepsilon_{X^{(j,\Delta f)}_{k,l}}] = \frac{1}{n^2}\frac{1}{l}\sum_{k=0}^{K-1}\sum_{l=0}^{L-1}\mathbb{E}[|X^{(\Delta f)}_{k,l}-\tilde{X}^{(z,\Delta f)}_{k,l}|^2].\tag{18}$$

From the (14) and (17), we can infer the following two conditions for successful channel estimation and information decoding:

1 if $\sigma^2_{q,n,l} > \sigma^2_{x,n,l}$, then interference from the information symbol would be lesser on the channel estimates.

2 if $\sigma^2_{x,n,l} > \sigma^2_{w,l}$, then interference from the channel noise would be lesser on the channel estimates and also on the information estimates.

Based on the above conditions, $\frac{\sigma^2_{\tilde{Q},n,l}}{\sigma^2_{x,n,l}}$ and $\frac{\sigma^2_{x,n,l}}{\sigma^2_{w,l}}$ are denoted as β_Q and β_X. The value of β_Q and β_X are used in Sec. IV to find the optimum error rate performance.

C. CHANNEL ESTIMATION AND INFORMATION DETECTION WITH IB-DFE

We employ an iterative receiver to improve the accuracy of the information estimates. This process is explained below

$$\tilde{X}^{(j,\Delta f)}_{k,l} = (Y^{(\Delta f)}_{k,l,i}-\alpha\sqrt{P_q}Q^{(\Delta f)}_{k,l}\tilde{H}^{(j)}_{k,l})F^{(j)}_{k,l}-\tilde{X}^{(j-1,\Delta f)}_{k,l}B^{(j)}_{k,l},\tag{19}$$

where $\tilde{X}^{(j-1,\Delta f)}_{k,l}$ is the previous iteration value of $\tilde{X}^{(j,\Delta f)}_{k,l}$ and j is the number of iteration followed in IB-DFE receiver and, $j = 0, 1, \ldots, J$. In the first iteration, the channel estimate obtained from (16) is used to estimate information in the IB-DFE receiver, then $\tilde{H}^{(0)}_{k,l} = \tilde{H}^{av}_{k,l}$. $F^{(j)}_{k,l}$ is the feed forward coefficient and it is written as

$$F^{(j)}_{k,l} = \frac{\tilde{F}^{(j)}_{k,l}}{\frac{1}{K}\sum_{k=0}^{K-1}(\tilde{F}^{(j)}_{k,l}\tilde{H}^{(j)}_{k,l})},\tag{20}$$

where $\tilde{F}^{(j)}_{k,l}$ is written as

$$\tilde{F}^{(j)}_{k,l} = \frac{\tilde{H}^{(j)}_{k,l}}{\left(\frac{\sigma^2_{n,k,l}}{\sigma^2_{x,n,l}}\right)+|\tilde{H}^{(j)}_{k,l}|^2\left(1-(\rho^{(j-1)})^2\right)},\tag{21}$$

and the correlation factor $\rho^{(j-1)} = \frac{\mathbb{E}[\hat{X}^{(j)}_{n,l}X^*_{n,l}]}{\mathbb{E}[|X_{n,l}|]}$. The feedback co-efficient $B^{(j)}_{k,l}$ is written as

$$B^{(j)}_{k,l} = F^{(j)}_{k,l}\tilde{H}^{(j)}_{k,l}-1.\tag{22}$$

The channel estimates can be further refined by using information estimates obtained from the iterative receiver. The expected value of information estimate error is calculated by using MSE is written as

$$\mathbb{E}[\varepsilon_{X^{(j,\Delta f)}_{k,l}}] = \frac{1}{n^2}\frac{1}{l}\sum_{k=0}^{K-1}\sum_{l=0}^{L-1}\mathbb{E}[|X^{(\Delta f)}_{k,l}-\tilde{X}^{(j,\Delta f)}_{k,l}|^2].\tag{23}$$

By using $\tilde{X}_{k,l}^{(j,\Delta f)}$ and $Q_{k,l}^{(\Delta f)}$ in (12), the new channel estimates can be written as

$$\tilde{H}_{k,l}^{(j)} = \alpha\left(\frac{H_{k,l}(\sqrt{P_x}\tilde{X}_{k,l}^{(j,\Delta f)} + \sqrt{P_q}\tilde{Q}_{k,l}^{(\Delta f)}) + W_l}{\sqrt{P_x}\tilde{X}_{k,l}^{(j,\Delta f)} + \sqrt{P_q}\tilde{Q}_{k,l}^{(\Delta f)}}\right) + l\frac{W_{e,l}}{\sqrt{P_x}\tilde{X}_{k,l}^{(j,\Delta f)} + \sqrt{P_q}\tilde{Q}_{k,l}^{(\Delta f)}}. \quad (24)$$

There are two channel estimates that are obtained from this receiver:

1 The channel estimation value obtained without using iterative receiver i.e. (16)
2 The channel estimation value obtained from the iterative receiver i.e. (24), this channel estimate improves with each IB-DFE iteration.

Applying (24) instead of (16) in (19) gives improved information estimates than the previous estimates, which is denoted as

$$\tilde{X}_{k,l,F}^{(j,\Delta f)}. \quad (25)$$

The expected value of information estimate error of $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ is calculated by using MSE is written as

$$\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}] = \frac{1}{n^2}\frac{1}{l}\sum_{k=0}^{K-1}\sum_{l=0}^{L-1}\mathbb{E}[|X_{k,l}^{(\Delta f)} - \tilde{X}_{k,l,F}^{(j,\Delta f)}|^2]. \quad (26)$$

There are three information estimates obtained from this receiver:

1 The information estimation value obtained by using average channel estimate (16) without the help of iterative receiver i.e. (17)
2 The information estimation value obtained by using average channel estimate (16) and also using iterative receiver i.e. (19)
3 The information estimation value obtained by using improved channel estimate obtained with the help of iterative receiver i.e. (24) and also by using iterative receiver i.e. $\tilde{X}_{k,l,F}^{(j,\Delta f)}$[1] as in (25)

The extrinsic information of $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ and $\tilde{X}_{k,l}^{(z,\Delta f)}$ are calculated by compensating the phase rotation on the estimated symbols present in $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ and $\tilde{X}_{k,l}^{(z,\Delta f)}$, respectively. Converting $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ and $\tilde{X}_{k,l}^{(z,\Delta f)}$ to time domain gives $e^{j\tilde{\theta}_n}\{\tilde{X}_{n,l,F}^{(j)}\} = \mathrm{IDFT}\{\tilde{X}_{k,l,F}^{(j,\Delta f)}; k = 0, 1, \ldots, K-1\}$, and $e^{j\tilde{\theta}_n}\{\tilde{X}_{n,l}^{(z)}\} = \mathrm{IDFT}\{\tilde{X}_{k,l}^{(z)}; k = 0, 1, \ldots, K-1\}$, respectively, where $n = 0, 1, \ldots, N-1$ and $l = 0, 1, \ldots, L-1$. Then, the extrinsic information of $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ and $\tilde{X}_{k,l}^{(z,\Delta f)}$ respectively, are written as

$$\begin{aligned}\widehat{X}_{n,l,F}^{(j)} &= e^{j\theta_n}e^{-j\tilde{\theta}_n}\{\tilde{X}_{n,l,F}^{(j)}\},\\ \widehat{X}_{n,l}^{(z)} &= e^{j\theta_n}e^{-j\tilde{\theta}_n}\{\tilde{X}_{n,l}^{(z)}\},\end{aligned} \quad (27)$$

where $n = 0, 1, \ldots, N-1$ and $l = 0, 1, \ldots, L-1$.

[1] Where F in the subscript denotes the final IB-DFE iteration.

IV. NUMERICAL RESULTS

In this section, we discuss the simulation setup and demonstrate the performance of the channel estimation method using IB-DFE receiver and SWIPT protocol. The bit error rate (BER) performance of the system is calculated by averaging BER of 1000 signal blocks, where $N = 256$ and $T_l = 1$ second. We assume $\eta_{EH} = 0.9$, $\chi = 2$, $\zeta_{SD} = 30$ dBm and $\Delta f = 0.2$ with distance $d_{SD} = 3$ m. The value of α is kept as 0.7, thus 70% of total transmit power is allocated for EH and 30% of power is allocated for ID. The power of superimposed signal is denoted as P_{si} and it is calculated as $P_{si} = 10^{\frac{P_x}{10}} + 10^{\frac{P_q}{10}}$. By using (2) and (3), the energy harvested from the received signal can be calculated. The CFO estimation error is denoted as $\varepsilon_{\Delta f}$, where $\varepsilon_{\Delta f} = \frac{|\Delta\tilde{f} - \Delta f|}{\Delta f}$. This section has 2 subsections and the subsections are explained as follows:

A Performance Analysis of CFO Estimation - in this section, the results of $\varepsilon_{\Delta f}$ based on Δf, L and P_q are listed in Tab. 1 and Tab. 2. Further the effect of $\varepsilon_{\Delta f}$ on the channel estimation and information decoding based on L and P_q is demonstrated in Fig. 3 and Fig. 4, respectively.
B Performance Analysis of Channel Estimation - in this section, the signal is considered to have no CFO (i.e. $\Delta f = 0$) to study the impact of channel estimation. The channel and estimation error based on the P_q are demonstrated in Fig. 5. BER performance of the signal based on L and P_q is demonstrated in Fig. 7 and Fig. 6, respectively. In Fig. 8, BER performance improvement due to IB-DFE is demonstrated. Further, Tab. 3 illustrates the result of EH at receiver and Fig. 9 demonstrates the optimum β_Q value for the received signal depending the noise power at the receiver.

TABLE 1. CFO estimate error ($\varepsilon_{\Delta f}$) based on L blocks are used for the estimation.

Δf \ L	2	3	5	7	9	10
0.10	0.0351	0.0067	0.0055	0.0045	0.0049	0.0035
0.15	0.0679	0.0126	0.0042	0.0053	0.0051	0.0043
0.20	0.0571	0.0081	0.0062	0.0047	0.0053	0.0038

TABLE 2. CFO estimate based on the power of the pilot signal is used for the estimation.

Δf \ P_q	17 dBm	19 dBm	21 dBm	23 dBm	25 dBm	27 dBm
0.10	0.0089	0.0084	0.0037	0.0042	0.0016	0.0018
0.15	0.0119	0.0078	0.0031	0.0016	0.0018	0.0014
0.20	0.0153	0.0090	0.0020	0.0018	0.0027	0.0021

A. PERFORMANCE ANALYSIS OF CFO ESTIMATION

The performance of CFO estimation method based on the number of blocks used in signal to find $\Delta\tilde{f}$ is demonstrated in Tab. 1 and Fig. 3.

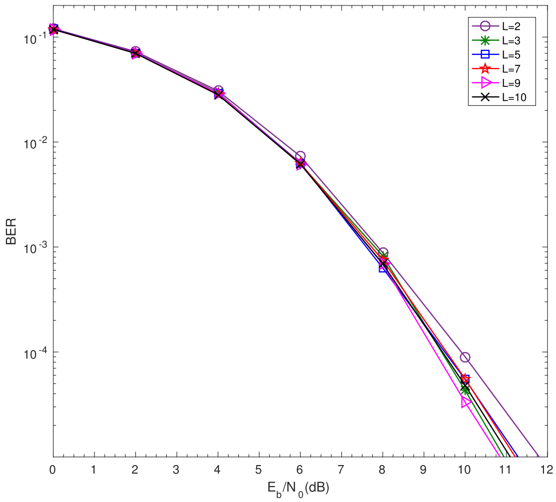

FIGURE 3. **Comparison of BER performance of system with CFO based on *L* blocks used for CFO estimation with constant P_q and P_x.**

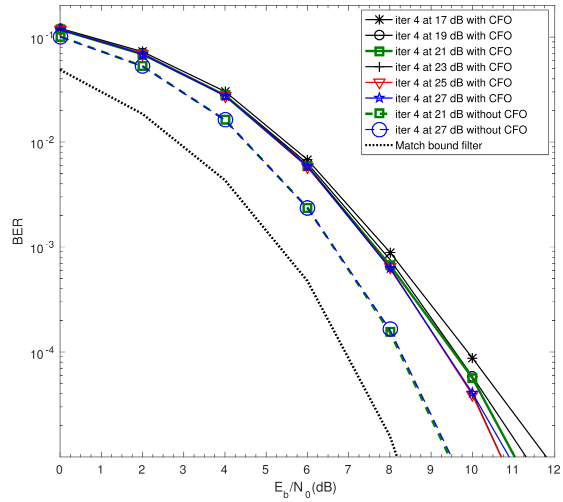

FIGURE 4. **Comparison of BER performance of system based on the power of the transmitted signal. Each curve demonstrate system performance based on P_q.**

In Tab. 1, $\varepsilon_{\Delta f}$ of the received signal is measured for various blocks with P_x and P_q are constant at 21 dBm and 25 dBm, respectively. The results show that at $\Delta f = 0.1$, $\varepsilon_{\Delta f}$ of the received signal with $\Delta f = 0.1$ has significant change between $L = 2$ to $L = 3$, but from $L = 3$ to $L = 10$, the changes are not significant and does not increase or decrease consistently. Similarly, the received signal with $\Delta f = 0.15$ and $\Delta f = 0.2$ has characteristics as $\Delta f = 0.1$. Therefore, it is better to fix $L = 3$ to improve estimation and keep the signal transmission more practical.

In Fig. 3, BER performance of the received signal is demonstrated with CFO estimation along with the channel and information estimation by using IB-DFE as in (24) and (25), respectively. P_x and P_q are constant at 21 dBm and 25 dBm, respectively and $\Delta f = 0.2$. The comparison of the signal with $L = 2$ over the signal with $L = 3$, illustrates that the BER improves significantly for the signal with $L = 3$ (i.e. SNR from 12 dB to 11 dB). But comparing the signal with $L = 3$ over the signal with $L > 3$, the improvement is relatively small by considering L blocks used (i.e. from 11 dB to 10.7 dB) and the relative improvement in BER performance with increase in every single block is inconsistent. The results of Fig. 3 is similar to Tab. 1 and it implies that the performance of the CFO estimation has a direct effect on channel and information estimation.

In Tab. 2, $\varepsilon_{\Delta f}$ of the received signal is measured for various P_q values with P_x is constant at 25 dBm and $L = 3$. The results illustrates that $\varepsilon_{\Delta f}$ of the received signal with $\Delta f = 0.1$ has significant change between $P_q = 17$ to 21 dBm, from $P_q = 21$ to 27 dBm, there are relatively small change and it is inconsistent. Similarly, the received signal with $\Delta f = 0.15$ and $\Delta f = 0.2$ has characteristics as $\Delta f = 0.1$.

In Fig. 4, BER performance of the received signal is demonstrated with CFO estimation along with the channel and information estimation by using IB-DFE as in (24) and (25), respectively. BER is simulated for signal with $P_q = 14$ to 25 and $P_x = 25$ dBm and $L = 3$ and $\Delta f = 0.2$. Similar to the results in Tab. 2, the signal with $P_q = 17$ dBm to 21 dBm has significant change in BER i.e. from 11.8 dB to 11 dB. BER performance for signal with $P_q = 21$ dBm to 25 dBm has only 0.3 dBm BER gain i.e. from 11 dB to 10.7 dB.

B. PERFORMANCE ANALYSIS OF CHANNEL ESTIMATION

Fig. 5 illustrates the optimum power to be used to transmit pilot signal when the information signal is transmitted at 25 dB and $L = 1$. The figure illustrates that curve A i.e. the expected value of channel estimate error drastically decreases approximately when pilot power is at 15 dBm to 17 dBm. After $P_q = 17$ dBm, the expected value decreases gradually with respect to increase in P_q. Curve B i.e the expected value

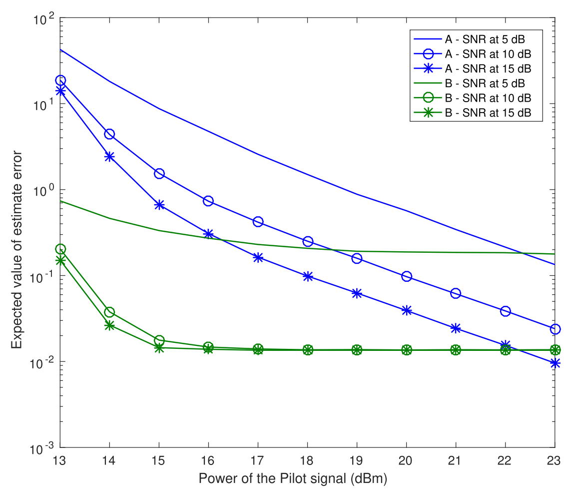

FIGURE 5. **The figure demonstrates results of estimate error versus the P_q at SNR 5 dB, 10 dB and 15 dB, respectively. The expected value of channel estimate error with iterative receiver is denoted as *A* and the expected value of information estimate error with IB-DFE receiver is denoted as *B*.**

of information estimate error decreases with increase in P_q and after $P_q = 17$ dBm, the curve B is showing constant gradient. A and B performance at high SNR is better than at low SNR. When P_x is at 25 dB, the optimum P_q is fixed 17 dBm since the curves of B at 10 dBm and 15 dBm SNR saturates when P_q is at 17 dBm.

Similar to Fig. 5, Fig. 6 gives a conclusion based on the BER performance of the system with increase in P_q. P_x is 25 dBm, P_q varies from at 14 dBm to 21 dBm and $L = 1$. The results demonstrates that if P_q is more than 17 dBm, then there is no significant improvement in BER performance. Fig. 7 demonstrates that increase in number of signal blocks to find the channel estimate average as in (16) improves the accuracy of the channel estimate and thereby improves the BER performance of the system. The signal blocks are averaged to improve the channel estimate accuracy by considering an assumption that the channel fading co-efficient is same for all the blocks. The figure also demonstrates that after using 2 slots, the channel estimate saturates to the ideal channel condition. Based on the results, we set $L = 1$ in Fig. 5, Fig. 6, Fig. 8, Fig. 9 and Tab. 3 to make the simulation more practical.

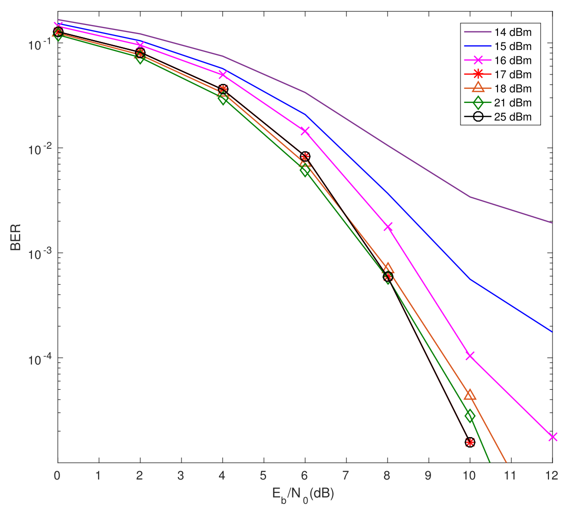

FIGURE 6. **Comparison of BER performance of system based on the power of the transmitted signal. Each curve demonstrate system performance based on P_q.**

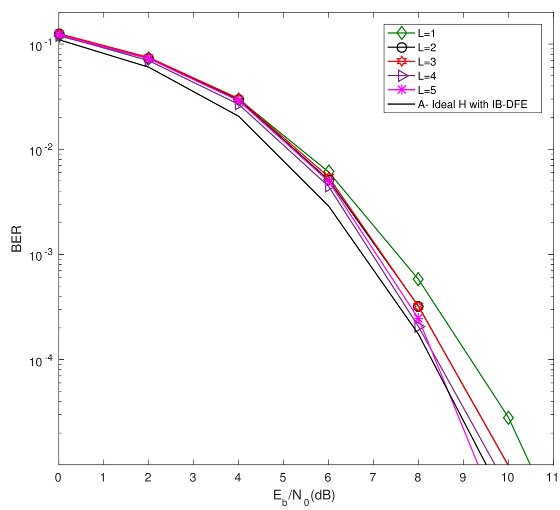

FIGURE 7. **Comparison of BER performance of system with iterative receiver based on the number of slots used to average the channel estimates. P_q and P_x is at 21 dBm and 25 dBm, respectively.**

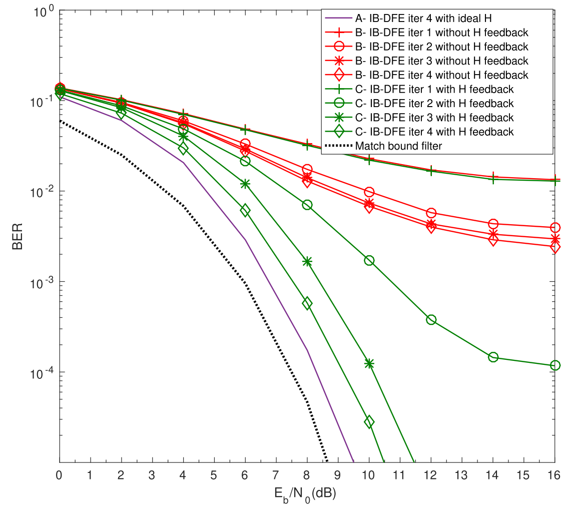

FIGURE 8. **Comparison of BER performance of system based on two different methods to estimate information.**

Fig. 8 demonstrates the BER performance of system based on two different methods to estimate information using iterative receiver. The curve A denotes the information estimate with the ideal channel condition. The information estimate obtained by using IB-DFE receiver only with information feedback is denoted as B. The information estimate obtained by using IB-DFE receiver with both channel feedback and information feedback is denoted as C. The information estimate found using match bound filter is denoted as D. P_x is 25 dBm, P_q is 21 dBm and $L = 1$. Fig. 8 demonstrates the improved BER performance of the system with IB-DFE receiver using both the channel and information estimate feedback. Due to the usage of both using channel estimate feedback and information estimate feedback recursively in the IB-DFE receiver, the BER performance of the information estimated as $\tilde{X}_{k,l,F}^{(j,\Delta f)}$ is better than the information estimated at (19). With each IB-DFE iteration, both the channel and information estimates improves, but this improvement ceases at 3^{rd} iteration compared to 4^{th} iteration. To validate our estimation techniques, the results are compared with the information estimate with the ideal channel and match bound filter.

Tab. 3 illustrates the value of P_{si}, E_y and $\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}]$ based on the value of P_q. $P_x = 25$ dBm and P_q varies from 14 dBm to 21 dBm. E_y increases proportionally with increase P_{si}. It is understood that the increase in P_q reduces $\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}]$ up to a certain limit. Since P_q is considerably lower than P_x, the percentage of increase in P_q value is higher as compared to percentage of increase in P_{si} value. Thereby, even with a slight increase in P_{si}, it is possible to improve the BER

TABLE 3. The amount of energy harvested at receiver and the expected value of information estimate error based on the power of pilot signal.

P_q (dBm)	14	15	16	17	18	19	20	21
P_{si} (dBm)	25.3320	25.4139	25.5150	25.6389	25.7901	25.9732	26.1933	26.4554
EH (mJ)	0.0212	0.0216	0.0221	0.0227	0.0235	0.0245	0.0258	0.0274
$\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}]$ at 5 dB SNR	0.4117	0.2756	0.2105	0.1815	0.1594	0.1559	0.1525	0.1411
$\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}]$ at 10 dB SNR	0.0356	0.0176	0.0143	0.0138	0.0138	0.0137	0.0138	0.0137
$\mathbb{E}[\varepsilon_{X_{k,l,F}^{(j,\Delta f)}}]$ at 15 dB SNR	0.0233	0.0136	0.0135	0.0135	0.0135	0.0135	0.0135	0.0135

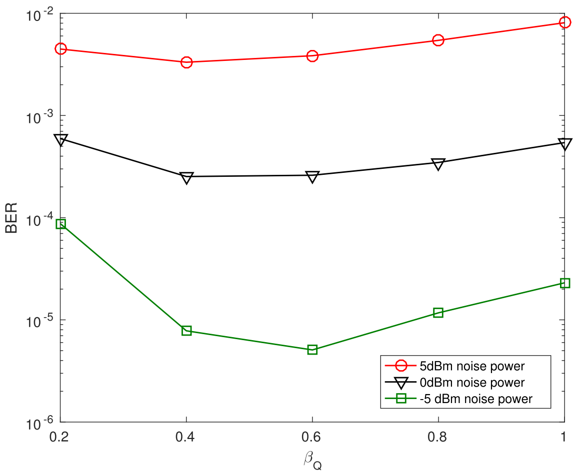

FIGURE 9. BER performance based on the ratio between the power of pilot and information in the superimposed signal.

performance drastically. Thus, this scheme is effective even at low SNR region.

Fig. 9 demonstrates the BER performance based on the ratio between the power of pilot and information in the superimposed signal and the optimum β_Q value for the received signal depending on the noise power at the receiver. This figure has different simulation setup, where $\Delta f = 0, L = 1$, $\chi = 2, d_{SD} = 3\,\text{m}, \alpha = 0.3, P_{si} = 28\,\text{dBm}$ with β_Q is varying from 0.2 to 1. There are three curves in the figure based on the signals with the fixed noise power -5 dBm, 0 dBm and 5 dBm, respectively at the receiver. The results prove that both the conditions $\sigma_{q,n,l}^2 > \sigma_{x,n,l}^2$ and $\sigma_{x,n,l}^2 > \sigma_{w,l}^2$ are necessary for better BER performance. The system with $\sigma_{w,l}^2 = 5$ has better BER performance at $\beta_Q = 0.4$ and the system with $\sigma_{w,l}^2 = -5$ has better BER performance at $\beta_Q = 0.6$. The curves explains that with the increase in noise power, the optimum β_Q reduces in order to allocate more power for the information signal and subsequently increasing β_X. Another important observation on β_Q by considering the curve with $\sigma_{w,l}^2 = -5$, the BER performance is better only if $\beta_Q = 0.6$ and not less than 0.6, otherwise channel estimation error will increase as compared to channel estimation error at $\beta_Q = 0.6$ and subsequently degrades the BER performance.

V. CONCLUSION

In this work, simultaneous wireless information and power transmission scheme is employed to harvest energy estimate CFO and channel condition by using superimposed pilot and information signal. We find the optimum ratio of the power required for the pilot signal and information signal to achieve a desirable error rate performance in respect to varying SNR conditions. Also, an algorithm is implemented at the receiver by using the feedback of the channel and information estimation to improve the performance. The presented analytical results are in line with the numerical results. This system can be extended to massive MIMO system model with wider acquisition range for the CFO.

REFERENCES

[1] D. N. K. Jayakody, J. Thompson, S. Chatzinotas, and S. Durrani, *Wireless Information and Power Transfer: A New Paradigm for Green Communications*. New York, NY, USA: Springer, 2017.

[2] T. D. P. Perera, D. N. K. Jayakody, S. K. Sharma, S. Chatzinotas, and J. Li, "Simultaneous wireless information and power transfer (SWIPT): Recent advances and future challenges," *IEEE Commun. Surveys Tuts.*, vol. 20, no. 1, pp. 264–302, 1st Quart., 2018.

[3] A. Rajaram, D. N. K. Jayakody, K. Srinivasan, B. Chen, and V. Sharma, "Opportunistic-harvesting: RF wireless power transfer scheme for multiple access relays system," *IEEE Access*, vol. 5, pp. 16084–16099, 2017.

[4] K. Raghunath and A. Chockalingam, "SC-FDMA versus OFDMA: Sensitivity to large carrier frequency and timing offsets on the uplink," in *Proc. IEEE Global Telecommun. Conf.*, Honolulu, HI, USA, Nov./Dec. 2009, pp. 1–6.

[5] M. Sabbaghian and D. Falconer, "Joint turbo frequency domain equalization and carrier synchronization," *IEEE Trans. Commun.*, vol. 7, no. 1, pp. 204–212, Jan. 2008.

[6] R. Dinis, A. Teresa, P. Pedro, and N. Fernando, "Joint turbo equalisation and carrier synchronisation for SC-FDE schemes," *Trans. Emerg. Telecommun. Technol.*, vol. 21, no. 2, pp. 131–141, 2010.

[7] F. Silva, R. Dinis, and P. Montezuma, "Frequency-domain receiver design for transmission through multipath channels with strong Doppler effects," *Wireless Pers. Commun.*, vol. 83, no. 2, pp. 1213–1228, Jul. 2015.

[8] H. Doğan, N. Odabasioglu, and B. Karakaya, "Time and frequency synchronization with channel estimation for SC-FDMA systems over time-varying channels," *Wireless Pers. Commun.*, vol. 96, no. 1, pp. 163–181, 2017.

[9] F. Silva, R. Dinis, and P. Montezuma, "Channel estimation and equalization for asynchronous single frequency networks," *IEEE Trans. Broadcast.*, vol. 60, no. 1, pp. 110–119, Mar. 2014.

[10] P. H. Moose, "A technique for orthogonal frequency division multiplexing frequency offset correction," *IEEE Trans. Commun.*, vol. 42, no. 10, pp. 2908–2914, Oct. 1994.

[11] T. M. Schmidl and D. C. Cox, "Robust frequency and timing synchronization for OFDM," *IEEE Trans. Commun.*, vol. 45, no. 12, pp. 1613–1621, Dec. 1997.

[12] M. Morelli and U. Mengali, "An improved frequency offset estimator for OFDM applications," *IEEE Commun. Lett.*, vol. 3, no. 3, pp. 75–77, Mar. 1999.

[13] N. Benvenuto and S. Tomasin, "Block iterative DFE for single carrier modulation," *Electron. Lett.*, vol. 38, no. 19, pp. 1144–1145, Sep. 2002.

[14] M. Jianpeng, Z. Shun, L. Hongyan, Z. Nan, and N. Arumugam, "Iterative LMMSE individual channel estimation over relay networks with multiple antennas," *IEEE Trans. Veh. Technol.*, vol. 67, no. 1, pp. 423–435, Jan. 2018.

[15] R. Dinis, C.-T. Lam, and D. Falconer, "Joint frequency-domain equalization and channel estimation using superimposed pilots," in *Proc. IEEE Wireless Commun. Netw. Conf.*, Mar./Apr. 2008, pp. 447–452.

[16] H. Sari, G. Karam, and I. Jeanclaude, "An analysis of orthogonal frequency-division multiplexing for mobile radio applications," in *Proc. IEEE Veh. Technol. Conf.*, Stockholm, Sweden, vol. 3, Jun. 1994, pp. 1635–1639.

[17] J. Rodríguez-Fernández, N. González-Prelcic, K. Venugopal, and R. W. Heath, "Frequency-domain compressive channel estimation for frequency-selective hybrid millimeter wave MIMO systems," *IEEE Trans. Wireless Commun.*, vol. 17, no. 5, pp. 2946–2960, May 2018.

[18] J. Guerreiro, R. Dinis, and P. Montezuma, "Analytical performance evaluation of precoding techniques for nonlinear massive MIMO systems with channel estimation errors," *IEEE Trans. Commun.*, vol. 66, no. 4, pp. 1440–1451, Apr. 2018.

[19] E. Dahlmaa *et al.*, "A framework for future radio access," in *Proc. IEEE 61st Veh. Technol. Conf. (VTC-Spring)*, vol. 5, May 2005, pp. 2944–2948.

AKASHKUMAR RAJARAM received the B.E. degree in electrical and electronics engineering from SKR Engineering College, Anna University, Chennai, India, in 2011, and the M.S. degree (engineering) in cyber security from the Tallinn University of Technology, Estonia, and the University of Tartu, Estonia, in 2016. He is currently pursuing the double Ph.D. degrees with the Universidade Nova de Lisboa, Portugal, and National Research Tomsk Polytechnic University, Russia. From 2017 to 2018, he was a Research Engineer with Tomsk Polytechnic University, Russia, and the Universidade Nova de Lisboa, Portugal. His research interests are in the areas of simultaneous wireless information and power transfer and channel estimation in wireless communication systems. He received the Best Thesis Award for the M.Sc. hardware and systems category from the ICT Thesis Contest 2016, Tallinn.

RUI DINIS received the Ph.D. degree from the Instituto Superior Técnico (IST), Technical University of Lisbon, Portugal, in 2001, and the Habilitation degree in telecommunications from the Faculdade de Ciências e Tecnologia (FCT), Universidade Nova de Lisboa (UNL), in 2010. He was a Researcher with the Centro de Análise e Processamento de Sinal, IST, from 1992 to 2005, and the Instituto de Sistemas e Robótica, from 2005 to 2008. From 2001 to 2008, he was a Professor with IST. In 2003, he was an invited Professor with Carleton University, Ottawa, Canada. Since 2009, he has been a Researcher with the Instituto de Telecomunicações. He is currently an Associate Professor with FCT, UNL. He is an IEEE Senior Member and an editor of the IEEE Transactions on Communications (transmission systems—frequency-domain processing and equalization), the IEEE Transactions on Vehicular Technology, *Physical Communication* (Elsevier), and *ISRN Communications and Networking* (Hindawi). He was also a Guest Editor of *Physical Communication* (Elsevier) (Special Issue on Broadband Single-Carrier Transmission Techniques) and is currently a Guest Editor of *Mobile Information Systems* (Hindawi) (Special Issue on Iterative Detection Schemes for MIMO Systems).

He has supervised 20 Ph.D. students (current and past) and has published 4 books, over 100 journal papers and book chapters and over 300 conference papers (of which 5 received best paper awards), and holds 10 patents (attributed or pending). He was involved in pioneer projects on the use of mm-waves for broadband wireless communications (international projects MBS and SAMBA), and his main research activities are on modulation and transmitter design and nonlinear effects on digital communications and receiver design (detection, equalization, channel estimation, and carrier synchronization), with an emphasis on frequency-domain implementations, namely for MIMO systems and/or OFDM and SC-FDE modulations. He is also involved in cross-layer design and optimization, involving PHY, MAC, and LLC issues, and indoor positioning techniques. He was part of the IEEE ICT'2014 Organizing Committee (TCP Co-Chair) and is or was a TPC Member for some of the major IEEE conferences, such as ICC, GLOBECOM, VTC, WCNC, and PIMRC. He is also a member of several technical committees of the IEEE Communications Society, such as SPCE, RCC, WC, and CT. He has been actively involved in several international research projects in the broadband wireless communications area (RACE project MBS, ACTS project SAMBA, and IST projects B-BONE and C-MOBILE) and many national projects, most of them as a nuclear researcher and/or in charge of his research center in multi-institutional projects.

DUSHANTHA NALIN K. JAYAKODY (S'09–M'14–SM'18) received the B.E. degree (Hons.) in electronics engineering in Pakistan (ranked as the merit position holder of the university under the SAARC Scholarship), the M.Sc. degree in electronics and communications engineering from the Department of Electrical and Electronics Engineering, Eastern Mediterranean University, Turkey (ranked as the first merit position holder of the department under the University Full Graduate Scholarship), and the Ph.D. degree in electronics, electrical, and communications engineering from University College Dublin, Ireland, in 2014. From 2014 to 2016, he was a Post-Doctoral Research Fellow with the Institute of Computer Science, University of Tartu, Estonia, and the Department of Informatics, University of Bergen, Norway. Since 2016, he has been a Professor with the School of Computer Science and Robotics, National Research Tomsk Polytechnic University, Russia, where he also serves as the Director of the Tomsk Infocomm Laboratory. He has received the Best Paper Award from the IEEE International Conference on Communication, Management and Information Technology, in 2017.

He has published over 80 international peer-reviewed journal and conference papers. His research interests include PHY layer prospective of 5G communications, cooperative wireless communications, device-to-device communications, LDPC codes, and unmanned aerial vehicle. He is a Senior Member of the IEEE. He has served as the Workshop Chair, the Session Chair, or the Technical Program Committee Member for various international conferences, such as the IEEE PIMRC 2013/2014, IEEE WCNC 2014–2018, and IEEE VTC 2015–2018. He currently serves as an Area Editor for the *Physical Communication Journal* (Elsevier), *Information* journal (MDPI), and the *Internet Technology Letters* (Wiley). Also, he serves as a reviewer for various IEEE Transactions and other journals.

NEERAJ KUMAR received the Ph.D. degree in CSE from SMVD University, Katra, India. He was a Post-Doctoral Research Fellow with Coventry University, Coventry, U.K. He is currently an Associate Professor with the Computer Science and Engineering Department, Thapar Institute of Engineering and Technology, Patiala. He has published more than 200 technical research papers in leading journals and conferences. He is an internationally renowned researcher in the areas of VANET, CPS smart grid, IoT mobile cloud computing, big data, and cryptography.