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Efficient equalisers for OFDM and DFrFT-OCDM multicarrier systems in mobile E-health video broadcasting with machine learning perspectives



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ABSTRACT

Recently, the orthogonal frequency-division multiplexing (OFDM) system has become an appropriate technique to be applied on the physical layer in various requests, mainly in wireless communication standards, which is the reason to use OFDM within mobile wireless medical applications. The OFDM with cyclic prefix (CP) can compensate lacks for the time-invariant multi-path channel effects using a single tap equaliser. However, for mobile wireless communication, such as the use of OFDM in ambulances, the Doppler shift is expected, which produces a doubly dispersive communication channel where a complex equaliser is needed. This paper proposes a low-complexity band LDL^H factorisation equaliser to be applied in mobile medical communication systems. Moreover, the discrete fractional Fourier transform (DFrFT) is used to improve the communication system's performance over the OFDM. The proposed low-complexity equaliser could improve the OFDM, and the DFrFT-orthogonal chirp-division multiplexing (DFrFT-OCDM) system's performance, as illustrated in the simulation results. This proves that the recommended system outperforms the standard benchmark system, which is an essential factor as it is to be applied within mobile medical systems.

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1. Introduction

Mobile wireless communication systems for E-health applications have received more attention recently with the goal to achieve a mobile hospital and patient monitoring system. Accordingly, the mobile wireless communication system, including video broadcasting features, is urgently needed in such applications. The orthogonal frequency-division multiplexing (OFDM) is the base for several communication systems such as, European digital video broadcasting systems like Digital Video Broadcasting-Terrestrial (DVB-T), DVB-Second Generation Terrestrial (DVB-T2),

DVB Handheld (DVB-H), Long-Term Evolution (LTE), and fifth generation (5G) mobile communication systems. The popularity of OFDM systems is based on its ability to compensate the effect for the time-invariant channel matrix. However, the OFDM loses its optimality against intercarrier interference (ICI) due to Doppler shift (doubly dispersive channel) or carrier frequency offset; accordingly, the system will be in need of sophisticated equalisers [1,2]. In [3,4], the discrete Fourier transform (DFT) was replaced by the discrete fractional Fourier transform (DFrFT) for multicarrier systems, which resulted in decreasing the Doppler frequency spread's effect, benefiting from the DFrFT subcarrier's chirped nature that mitigates the Doppler shift. As such, the ICI was reduced.

While DFrFT gives a more improved performance than DFT under the doubly dispersive fading channel, there is still a need for a complex equaliser [5]. Simple equalisers were proposed in [6–8], wherein [6] least-squares problems (LSQR) algorithm was offered to solve the matrix inversion iteratively; accordingly,

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no matrix inversion is needed, which simplifies the equaliser. In [7], the equaliser was simplified by using a banded matrix, while in [8], a new approach is proposed, which is based on both a banded matrix and the LDL^H factorisation algorithm.

Unlike the aforementioned simple equalisers which were applied with OFDM, this paper proposes the Orthogonal Chirp Division Multiplexing (DFrFT-OCDFM) systems, and then combines the simple suggested equaliser under a time-variant multi-path channel, which is deemed to be suitable for a medical mobile video broadcasting system. Furthermore, the DFrFT-OCDFM system was introduced in detail, including the way it is able to replace the DFT on the OFDM, and primarily, the simple equaliser has been added with DFrFT-OCDFM to fit the mobile medical applications. Moreover, the doubly dispersive channel details, with their effects on the OFDM and the DFrFT-OCDFM system's performance, will be outlined. The equalisation challenges will be specified, then an assessment between some known complex equalisers will be delivered to evaluate the equalisers' behaviour when improving the systems. The suggested simple equalisation methods based on the LDL^H factorisation algorithm is explained and presented as a practical solution for mobile medical applications.

802.11-WLAN video streaming was investigated in [9] over m-health claims, where a medical channel-adaptive fair allocation scheme was proposed to enhance the Quality of service (QoS) for IEEE 802.11 (WLAN). More recent work against real-time medical applications were explained in [10], where an adaptive video encoder compared to a real-time medical use is investigated to maximise the encoded video's quality, improve encoding rate, and to minimise the bit rate demands. In [11], an experimental set was introduced to provide mobile WiMAX video streaming performance analysis for Bandwidth on demand (BOND) services. More recent research and proposed wireless medical applications can be found in [12–15]. Comprehensive knowledge regarding the structure of health monitoring and machine learning can be found in the well-cited reference book [16], where the theory and the demonstration of the health monitoring structure were presented. Recently, a lot of research has been carried out in this field, including, [17], where the limitations of machine learning approaches have been investigated, and future clinical translations defined. Specific application for using machine learning within health monitoring is rapidly increasing, for example, in [18] where this technique was proposed for the early prediction of asthma attacks.

On the other hand, [19] investigates the scenario of E-health applications that apply the multi-service stream network, which concludes that the mathematical model class $G/G/1$ – in its general case of a single-channel system – is regarded as an appropriate technique to be implemented within the E-health applications. Indeed, the short time delay and the jitter are practically suitable for E-health primarily. Moreover, the packet losses and the error rates are also considered to be suitable within E-health.

The paper is organised as follows: In Section 2, the background of the OFDM system equalisation is explained to equip the reader with a more comprehensive understanding of the research work presented. The preliminaries for this research are also stated in this section. The proposed approach is presented in Section 3. Section 4 details the performance analysis of the proposed method, technical discussion, and deep computing/machine learning perspectives. Finally, conclusions are presented in the last section.

2. Background and preliminaries

The OFDM allows high data rates to be reliably and efficiently transmitted over the delay-dispersive channels. By dividing the transmitted signal into several narrow bandwidth sub-carriers,

OFDM can mitigate the undesirable multi-path effects, mainly, the inter symbol interference (ISI) quantity in long symbol time systems. Moreover, at the beginning of each symbol, a guard period is added – termed cyclic prefix (CP) – to eliminate the expected effects of ISI over the multi-path signals' delay. The multi-path effect tolerant resulting from CP makes OFDM more suitable for high data rate transmission over terrestrial locations rather than single carrier transmissions.

The CP has significant influence over the OFDM system equalisation, as a result of inserting it at the first part of the OFDM symbol. This transfers the multi-path frequency fading channel matrix into a circulant matrix that can be diagonalised by the FFT at the receiver side. The diagonalised channel matrix can be compensated using a single tap equaliser, which can be considered as a simple multiplication in the gain and phase components.

The basic block diagram for a baseband OFDM transmission and reception system is shown in Fig. 1. The OFDM signal is corrupted by passing through the channel. Taking into consideration that the receiver's mission is to obtain the useful information from the corrupted message correctly, accordingly, the receiver converts the received signal into its original form depending on a single tap equaliser.

Let us consider the communication system in Fig. 1 in its sequence processing steps over a noisy, frequency fading channel. The received symbols are shown in (1):

$$\mathbf{z}_n = \mathbf{H}\mathbf{F}^H \mathbf{x}_n + \mathbf{v}_n \quad (1)$$

where \mathbf{z}_n is the received signal, \mathbf{H} is the $N \times N$ frequency fading channel matrix, N is the number of subcarriers, \mathbf{F}^H is the inverse discrete Fourier transform (IDFT) matrix of DFT, \mathbf{x}_n is the data vector transmitted in the n th OFDM symbol, and \mathbf{v}_n is the time domain of the white Gaussian noise (WGN). After demodulation and using DFT, the received vector can be calculated as:

$$\tilde{\mathbf{z}}_n = \mathbf{F}\mathbf{H}\mathbf{F}^H \mathbf{x}_n + \mathbf{F}\mathbf{v}_n \quad (2)$$

Where \mathbf{H} is a circulant matrix (resulting from CP), $\mathbf{F}\mathbf{H}\mathbf{F}^H$ becomes a diagonal matrix [20], and we can equalise the received signal by simple adjustment of the amplitude and phase for the received sequence [21]. This property is one of the key advantages of OFDM as it reduces the complexity of the equalisation process in a multi-path fading channel, which is a harsh environment requiring complex equalisers [22]. However, this property is valid only in time-invariant frequency-selective multi-path channels [23].

When the channel is doubly selective or the receiver induces a frequency offset, [24], the channel matrix is no longer circulant, the DFT cannot diagonalise and ICI appears. In this event, OFDM needs a complicated equaliser [8,22,25], such as the minimum mean square error (MMSE) equaliser. The fractional Fourier transform (FrFT) proposed a new base for OFDM [3,4] that can enhance multicarrier modulation (MCM) systems' performance under a doubly dispersive fading channel because of its ability to cope with the Doppler shifts and to compensate its effects.

2.1. Discrete Fractional Fourier Transform (DFrFT)

The FrFT was presented as a new idea in 1929 [26] as a generalisation of the FT. Namias reintroduced FrFT in mathematics for applications in quantum mechanics in 1980 [27]. The DFrFT appeared after many groups of researchers reinvented FrFT [28]. Later, low-complexity representations, computational cost, and applications for the DFrFT were investigated in [29,30]. Nowadays, DFrFT is being used in various requests, such as in optics, image processing, and signal processing.

One of the FrFT definitions is that:

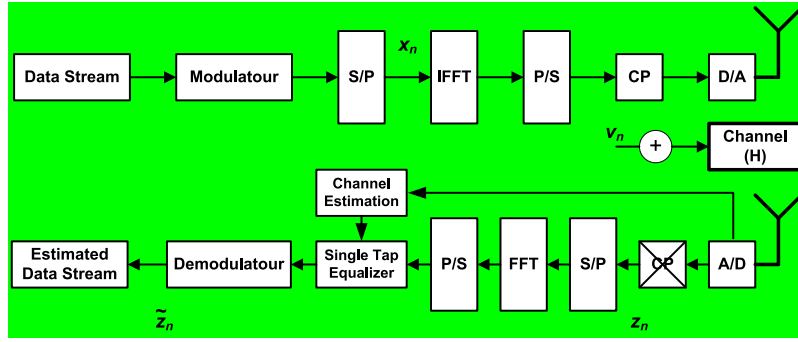


Fig. 1. Basic baseband OFDM transmission and reception system.

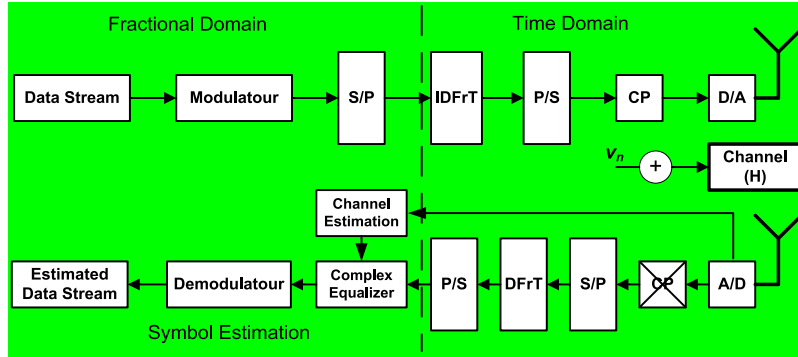


Fig. 2. The Basic baseband DFrFT-MCM transmission and reception system with a complicated equaliser.

“A fractional Fourier transform is a rotation operation on the time–frequency distribution by angle” [28].

For $\alpha = 0$ when DFrFT has no effect, for $\alpha = \pi/2$ when DFrFT returns to FT, and for any value of α in between 0 to $\pi/2$; the DFrFT substitutes the time–frequency distribution based on the value of α .

The transformation kernel of the continuous FrFT is according to [30]:

$$K_{\alpha}(t, y) = A_{\alpha} e^{j\pi(t^2+y^2)\cot\alpha - j2\pi ty \csc\alpha} \quad (3)$$

where α is the rotation angle for the transformation process and

$$A_{\alpha} = \frac{e^{[-j\pi \text{sign}|\sin\alpha|/4 + j\alpha/2]}}{\sqrt{|\sin\alpha|}} \quad (4)$$

The FrFT becomes:

$$f_{\alpha}\{d(t)\}(y) = X_{\alpha}(y) = \int_{-\infty}^{\infty} d(t) K_{\alpha}(t, y) dt \quad (5)$$

$$d(t) = \int_{-\infty}^{\infty} D_{\alpha}(y) K_{-\alpha}(t, y) dy \quad (6)$$

The fractional Fourier signal domain is defined by the α angle for $0 < |\alpha| < \pi$. Fourier transform can be obtained using $\alpha = \pi/2$. There are several DFrFT algorithms with various properties and accuracies. The DFrFT algorithm proposed in [31] is used in this work. Suppose that the input and output functions of the DFrFT $f(t)$ and $F_{\alpha}(y)$ respectively have the chirp period of order p with the period $T_p = N\Delta t$, $F_p = M\Delta y$, and the sampled signals are bound between the interval Δt and Δy as:

$$d(n) = f(n\Delta t), D_{\alpha}(m) = F_{\alpha}(m\Delta y) \quad (7)$$

where $n = 0, 1 \dots N-1$ and $m = 0, 1 \dots M-1$. When $\alpha \neq X.\pi$ (X is an integer), (5) can be converted to:

$$D_{\alpha}(m) = A_{\alpha} \Delta t e^{\frac{j}{2} \cot\alpha m^2 \Delta y^2} \sum_{n=0}^{N-1} e^{\frac{j}{2} \cot\alpha n^2 \Delta t^2} e^{j \csc\alpha n.m.\Delta t.\Delta y} d(n) \quad (8)$$

when $M = N$ the transformation is reversible, with the condition:

$$\Delta t \Delta y = 2\pi \sin\alpha / M \quad (9)$$

Eq. (8) may also be written in a multiplication of matrix and vector form,

$$\mathbf{D} = \mathbf{F}_{\alpha} \mathbf{d} \quad (10)$$

where $\mathbf{D} = [D_{\alpha}(0), D_{\alpha}(1), \dots, D_{\alpha}(N-1)]^T$, $\mathbf{d} = [d(0), d(1), \dots, d(N-1)]^T$, and \mathbf{F}_{α} is an $N \times N$ matrix in the same way, the IDFrFT may be written as:

$$\mathbf{d} = \mathbf{F}_{-\alpha} \mathbf{D} \quad (11)$$

where $F_{-\alpha} = F_{\alpha}^H$.

The DFrFT-MCM system shown in Fig. 2 is a system based on block data transfer, and the subcarriers are orthogonal to each other where each subcarrier is a different chirp signal, so we can call it DFrFT-OCDM. Two bases for the DFrFT-OCDM system are shown in Fig. 3 and the spectral energy distribution for the two bases are shown in Fig. 4.

The Wigner distribution in time and frequency domain for the 1st basis signal and the 20th basis signal with $\alpha = 0.7$ are shown in Fig. 5. The figure shows that the DFrFT bases are frequency varying with time, which is a property of the DFrFT transformation.

The transmitted DFrFT-OCDM signal is a combination of many blocks, starting with a CP to eliminate the ISI. However, the need for fast mobile communications with significant high data rates

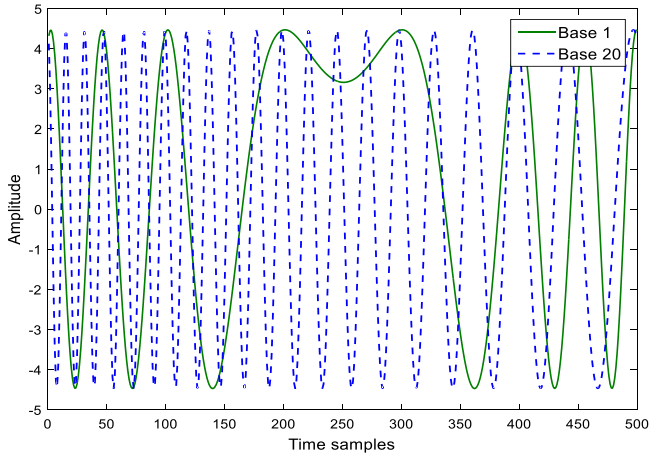


Fig. 3. DFrFT-OCDM basis for the 1st and the 20th basis signals.

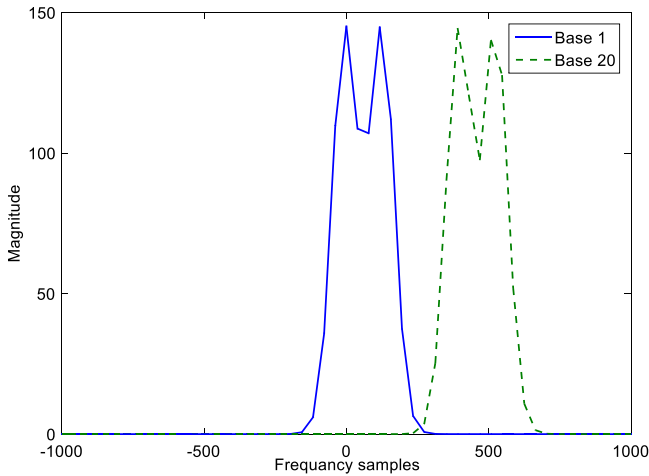


Fig. 4. Spectral Energy Distribution for the 1st and the 20th basis signals.

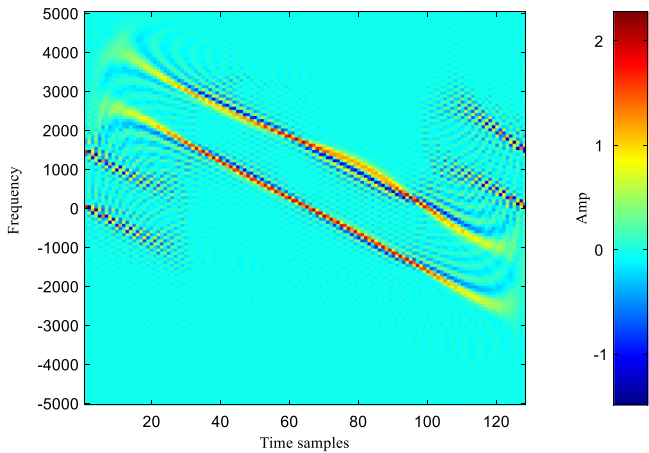


Fig. 5. The Wigner distribution for the 1st basis signal and the 20th basis signal.

and long symbols introduces larger ICI, which forces the use of complicated equalisers.

The DFrFT-OCDM system complexity is nearly equivalent to the traditional OFDM system [3], and both systems exhibit almost the same performance when the channel is time-invariant. However, neither of them can diagonalise the time-variant channel

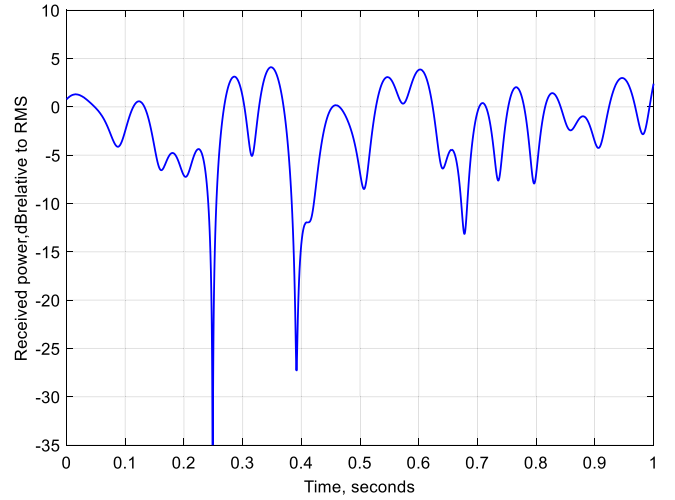


Fig. 6. Rayleigh fading channel corresponding to a Doppler shift of 10 Hz.

matrix; the DFrFT-OCDM can compress it towards the diagonal much more effectively than OFDM, which is the main advantage of the DFrFT-OCDM system. This enables it to achieve a better performance than OFDM and provides the opportunity to use low-complexity equalisers while maintaining this better performance.

2.2. Doubly dispersive channel

The channels of the mobile-radio applications have time-variant behaviour, due to the transmitter and/or the receiver movement that results in the continual changing of the propagation paths. The changing pace of the propagation circumstances is causally related to the fading rapidity, i.e. the speed of the changing rate of fading environments.

In the environment of proposed multimedia mobile communications for medical applications, multiple copies of the transmitted signal are received with different delays and phases at the receiver. This creates the phenomenon of multi-path, resulting in a random frequency modulation on each of its multi-path components due to the Doppler shifts. Hence, the resultant received signal may suffer from severe attenuation and interference that can lead to errors at the receiver and system performance degradation.

The variation in the fading channel frequency response with time due to the Doppler shift in fading channels is called a doubly dispersive fading channel. The Doppler shift calculations are given by:

$$f_d = (\Delta u/C)f_c \quad (12)$$

where f_d is the Doppler shift frequency, Δu is the velocity difference between the transmitter and the receiver, C is the speed of light, and f_c is the signal carrier frequency.

Figs. 6 and 7 show the Rayleigh fading channel mutual shape for two different maximum Doppler shift frequencies of 10 Hz and 100 Hz, respectively. These Doppler shifts are measured respectively for 6 km/h and 60 km/h velocities at 1800 MHz, which is one of the working frequencies for GSM mobile networks. The intensity of the signal can fall by several thousand factors or 30–40 dB in some “deep fades”.

2.3. Communication channel equalisation

The majority of errors at the receiver side are caused by the channel distortion, whilst the most effective method to compensate for this channel distortion effect, in order to recover the

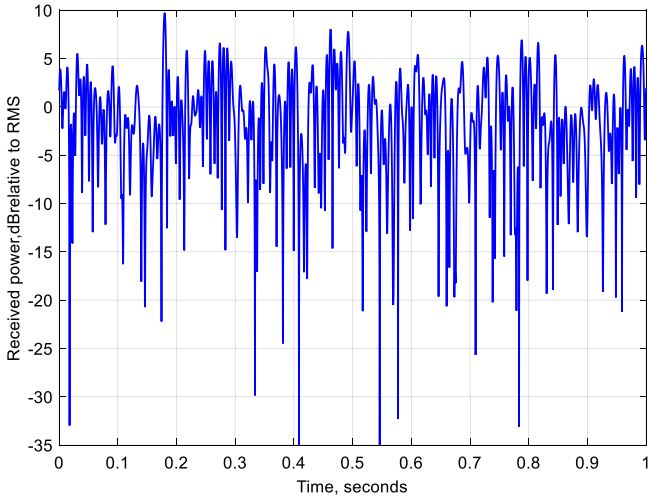


Fig. 7. Rayleigh fading channel corresponding to a Doppler shift of 100 Hz.

original signal's shape, is the equalisation, as shown, for example, in Fig. 8 [32]. The most fundamental method used by the equalisation is to select the correct receiver's filter to compensate for the selectivity of the radio channel frequency completely. This could be accomplished by choosing the receiver's filter impulse response that satisfies the relation in (13):

$$\mathbf{W} \otimes h = 1 \quad (13)$$

where \mathbf{W} is the equaliser impulse response, h is the channel impulse response, and " \otimes " represents the linear convolution. This algorithm of equalisation is called "zero-forcing (ZF) equalisation" and, according to [32–34], ZF can provide the complete removal of any frequency selectivity in the radio channel. As a result, destruction-free and corruption-free signals can be achieved. However, ZF equalisation may become a great source of noise amplification that occurs after the filtering process. This may have an undesirable effect in that it may cause severe degradation in the system's performance.

An alternative suggestion for the ZF equalisation is to build a filter that provides a compromise between the noise/interference level and the signal distortion level based on the level of the radio-channel frequency selectivity. This might be accomplished by the MMSE equaliser that selects the filter to minimise the mean-square error (ε) between the transmitted signal and the equaliser output:

$$\varepsilon = E \left\{ \left| \hat{z}(t) - z(t) \right|^2 \right\} \quad (14)$$

where $\hat{z}(t)$ is the estimated signal, and $z(t)$ is the actual transmitted signal.

It was proved that using a single carrier modulation system in frequency fading channel is inefficient due to the time equalisation complexity. On the other hand, the OFDM systems give an instant solution to this problem using a single tap equaliser in the frequency domain. As the OFDM cannot deal with doubly dispersive channels, there is a motivation to search for other bases that can match the channel frequency variation with time like DFrFT.

3. Proposed method

Fig. 9 shows the OFDM system data flow, $\mathbf{x}_n = [x_0, x_1, \dots, x_{N_a-1}]^T$ is the data vector transmitted in the n th OFDM symbol, whilst its samples are permuted by the binary matrix $\mathbf{P} \in \mathbb{Z}^{N \times N_a}$ in the

frequency domain, which allocates a data vector $\mathbf{x}_n \in \mathbb{C}^{N_a}$ to N subcarriers, with only N_a active:

$$\mathbf{P} = \left[\mathbf{0}_{N_a \times (N-N_a)/2} \mathbf{I}_{N_a} \mathbf{0}_{N_a \times (N-N_a)/2} \right] \quad (15)$$

\mathbf{I}_{N_a} is an identity matrix with $N_a \times N_a$ dimensions. The vector $\mathbf{s}_n = [s_0 s_1 \dots s_N]^T$ is calculated from:

$$\mathbf{s}_n = \mathbf{F}^H \mathbf{P} \mathbf{x}_n \quad (16)$$

where \mathbf{F}^H is the N -point IDFT matrix.

The time and frequency fading channel can be demonstrated by the time-variant discrete impulse response: $h(n, u)$, where n is the time instant, and u is the time delay. The justification of this model with further details can be found in [1,35,36] that could be stated in the formula of (time-variant, or circular) convolution matrix by:

$$[H]_{n,u} := h(n, (n-u)_N) \quad (17)$$

Supposing the causal channel and the maximum delay spread N_h were shorter than the CP $N_h \leq L$; after removing the CP, the n th OFDM received symbol can be specified by:

$$\mathbf{z}_n = \mathbf{H}_n \mathbf{x}_n + \mathbf{v}_n \quad (18)$$

where \mathbf{v}_n are the samples of the additive white Gaussian noise (AWGN) with variance of σ^2 . In static setting, \mathbf{H}_n is circulant and the DFT matrix can be decoupled. The received subcarriers are demodulated by DFT:

$$\mathbf{y} = \mathbf{F} \mathbf{z}_n \quad (19)$$

where \mathbf{F} is the DFT matrix. The equaliser matrix $\mathbf{W}_n \in \mathbb{C}^{N_a \times N_a}$ deals with the input:

$$\tilde{\mathbf{z}}_n = \mathbf{P}^H \mathbf{F} \mathbf{H}_n \mathbf{F}^H \mathbf{P} \mathbf{x}_n + \mathbf{P}^H \mathbf{F} \mathbf{v} = \mathbf{U}_n \mathbf{x}_n + \tilde{\mathbf{v}}_n \quad (20)$$

with a system matrix $\mathbf{U}_n \in \mathbb{C}^{N_a \times N_a}$, where $\mathbf{U}_n = \mathbf{P}^H \mathbf{F} \mathbf{H}_n \mathbf{F}^H \mathbf{P}$. \mathbf{P} is a binary matrix used to remove the components that may appear in the lower left and upper right corners in \mathbf{U}_n [37], and to support reducing the out-of-band emissions. The estimated data vector after using equaliser is specified by:

$$\hat{\mathbf{x}}_n = \mathbf{W} \tilde{\mathbf{z}}_n \quad (21)$$

It is easy to show that $[\tilde{H}]_{m,k} = \tilde{h}(m-k, k)$, where

$$\tilde{h}(m, k) = \frac{1}{N} \sum_{n=0}^{N-1} \sum_{u=0}^{N-1} h(n, u) e^{-j2\pi(uk+mn)/N} \quad (22)$$

From (22), it is clear that $\{\tilde{h}(0, :)\}$ is on the main diagonal of $[\tilde{H}]_{m,k}$, $\{\tilde{h}(1, :)\}$ and $\{\tilde{h}(-1, :)\}$ is on the first sub-diagonal and the first super-diagonal respectively. It is obvious that $\tilde{h}(m, k)$ is the response of the frequency-domain, at subcarrier $k+m$, to a frequency-domain impulse centred at subcarrier k . In $\tilde{h}(m, k)$, k is the frequency index and m known as Doppler index. In $h(n, u)$, n is known as the time index and u as the lag index.

The DFrFT-OCDFM system data flow diagram is shown in Fig. 10, which illustrates the difference from the OFDM system by using inverse fractional Fourier transform (IDFrFT) and the DFrFT for modulation and demodulation respectively. Using identical sequences for the data vector in the transmitter and the receiver, it can be shown that the equaliser $\mathbf{W}_n \in \mathbb{C}^{N_a \times N_a}$ function is to estimate the transmitted data using Eq. (20):

$$\tilde{\mathbf{z}}_n = \mathbf{P}^H \mathbf{F}_\alpha \mathbf{H}_n \mathbf{F}_{-\alpha} \mathbf{P} \mathbf{x}_n + \mathbf{P}^H \mathbf{F}_\alpha \mathbf{v} = \mathbf{U}_{n,\alpha} \mathbf{x}_n + \tilde{\mathbf{v}}_n \quad (23)$$

where \mathbf{F}_α and $\mathbf{F}_{-\alpha}$ represent the DFrFT matrix and the IDFrFT matrix respectively, with fractional angle $= \alpha$. The channel matrix

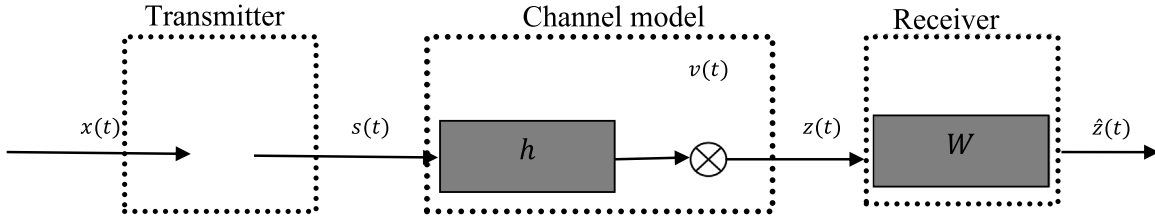


Fig. 8. General time-domain linear equaliser.

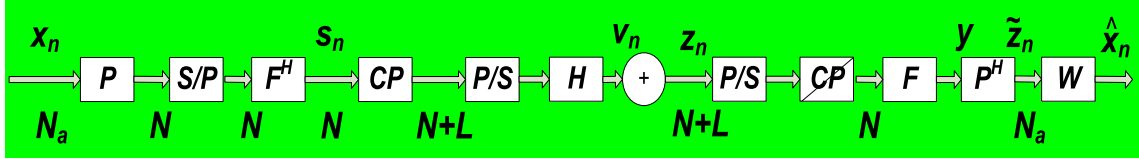


Fig. 9. OFDM data flow block diagram.

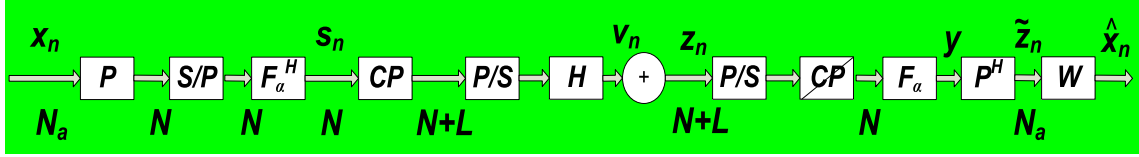


Fig. 10. DFrFT-OCDM System data flow block diagram.

and the noise vector in the fractional domain are given by $\tilde{\mathbf{H}}_\alpha = \mathbf{F}_\alpha \mathbf{H} \mathbf{F}_\alpha^{-1}$ and $\tilde{\mathbf{v}} = \mathbf{F}_\alpha \mathbf{v}$, respectively.

$\tilde{\mathbf{H}}$ and $\tilde{\mathbf{H}}_\alpha$ introduce ICI because they are non-diagonal sub-carrier channel matrices, which is the case in a doubly dispersive fading channel, accordingly, the process of the symbol estimation will be complicated and as a result, it is necessary to use a complex equaliser.

3.1. Zero forcing and MMSE block equalisers

The ZF and MMSE equalisers can estimate the transmitted data by minimising $E \{ \|\mathbf{x}_n - \mathbf{W}\tilde{\mathbf{z}}_n\|^2 \}$ [37]:

$$\hat{\mathbf{x}}_{ZF} = \tilde{\mathbf{H}}_\alpha^+ \tilde{\mathbf{z}}_n = \tilde{\mathbf{H}}_\alpha^H (\tilde{\mathbf{H}}_\alpha \tilde{\mathbf{H}}_\alpha^H)^{-1} \tilde{\mathbf{z}}_n \quad (24)$$

$$\hat{\mathbf{x}}_{MMSE} = \tilde{\mathbf{H}}_\alpha^H (\tilde{\mathbf{H}}_\alpha \tilde{\mathbf{H}}_\alpha^H + \gamma^{-1} \mathbf{I}_{N_a})^{-1} \tilde{\mathbf{z}}_n \quad (25)$$

when $\alpha = \pi/2$, the fractional domain channel matrix $\tilde{\mathbf{H}}_\alpha$ can be reduced to the frequency domain channel matrix $\tilde{\mathbf{H}}$, $\hat{\mathbf{x}}_{ZF}$, and $\hat{\mathbf{x}}_{MMSE}$ is the estimated data using the ZF and the MMSE equalisers respectively, γ is the signal-to-noise ratio (SNR), and $\tilde{\mathbf{H}}_\alpha^+$ is the fractional domain Moore-Penrose pseudo-inverse of the channel matrix [38]. In (24) and (25), complete information of the channel matrix \mathbf{H}_α is presumed thanks to the channel estimation, even when the guard subcarriers are not used by the equaliser. Moreover, it is presumed that: $E \{ \mathbf{x}_n \} = E \{ \tilde{\mathbf{v}}_n \} = 0$, $E \{ \mathbf{x}_n \mathbf{x}_n^H \} = \mathbf{I}$, $E \{ \mathbf{d}_n \tilde{\mathbf{v}}_n^H \} = 0$, and $E \{ \tilde{\mathbf{v}}_n \tilde{\mathbf{v}}_n^H \} = \sigma^2 \mathbf{I}$

The ZF equaliser enhances the noise so its performance is poor, whilst the performance of the MMSE equaliser is the best in all linear equalisers [5]. However, it is the most complicated because it needs channel matrix inversion that involves $\mathcal{O}(N_a^3)$ complex processes [39]. For high values of N_a like DVB-T, DVB-H and WiMAX, it is not practical.

3.2. Main contribution

Reduced complexity MMSE equalisers are proposed in [8, 22, 25, 37, 40–43]. In [37], a sequential MMSE equaliser is suggested and banded equalisers were presented in [8]. As identified in [37], a nearly-banded channel matrix produced in the frequency and fractional domains under doubly dispersive channels based on these conditions adapting LDL^H factorisation, can reduce the MMSE equaliser complexity [8, 20]. All the low-complexity equalisers are itemised for the OFDM systems alone; this is not the case for the DFrFT-OCDM systems.

LDL^H factorisation equaliser (Linear equaliser) was proposed in [42] as a low-complexity equaliser for the OFDM systems, benefiting from the banded properties of the frequency domain channel matrix $\tilde{\mathbf{H}}$. In like manner, the DFrFT-OCDM systems can use the LDL^H factorisation equaliser because the system matrix in the fractional domain is almost banded more than the system matrix in the frequency domain [3].

The calculation of the equaliser matrix \mathbf{W}_n is restricted to the first Q sub- and super-diagonals of $\tilde{\mathbf{H}}_\alpha$ by applying the binary masking matrix \mathbf{M} with elements:

$$\mathbf{M}(m, n) = \begin{cases} 1 & 0 \leq |m - n| \leq Q \\ 0 & Q < |m - n| < N_a \end{cases} \quad (26)$$

where the masked matrix:

$$\mathbf{B}_n = \mathbf{M} \odot \tilde{\mathbf{H}}_\alpha \quad (27)$$

which is shown in Fig. 11, where \odot denotes to the element-wise multiplication. The MMSE equaliser can be defined according to [42] as:

$$\mathbf{W}_{n,MMSE} = \mathbf{B}_n^H (\mathbf{B}_n \mathbf{B}_n^H + \gamma^{-1} \mathbf{I}_{N_a})^{-1} \quad (28)$$

where \mathbf{B}_n is a banded matrix with Q off-diagonal terms below and above the diagonal, which corresponds to a band structure for $(\mathbf{B}_n \mathbf{B}_n^H)$, only the first $2Q$ off-diagonal terms above and below the diagonal enclosed elements can reduce the calculation complexity of the MMSE equaliser in (28). In the meantime, (28)

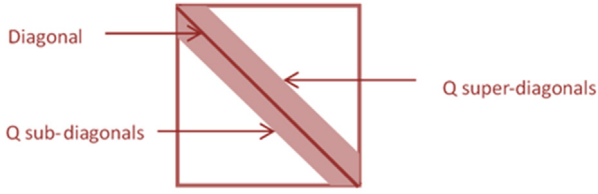


Fig. 11. The desired structure of the band matrix B inside the whole matrix \tilde{H}_α .

is regarded as time-dependent where $\hat{\mathbf{x}}_n = \mathbf{W}_{n,MMSE}\tilde{\mathbf{z}}_n$ can be deduced with no need for explicitly determining $\mathbf{W}_{n,MMSE}$.

The LDL^H factorisation of the Hermitian band matrix $\mathbf{B}_n\mathbf{B}_n^H + \gamma^{-1}\mathbf{I}_{N_a} = \mathbf{LDL}^H$ can then be directly calculated [39], reaching to:

$$\hat{\mathbf{x}}_n = \mathbf{B}_n^H (\mathbf{LDL}^H)^{-1} \tilde{\mathbf{z}}_n = \mathbf{B}_n^H \mathbf{d}_n \quad (29)$$

As an alternative option to calculate the inverse in (29), the system can be resolved by forwarding the substitution to obtain $\mathbf{d}_{2,n}$ via the lower left triangular matrix \mathbf{L} and a rescaling by the diagonal matrix \mathbf{D}^{-1} to calculate $\mathbf{d}_{1,n}$. Finally, back substitution with the upper right triangular \mathbf{L}^H yields \mathbf{x}_n , which can be inserted into (29) to determine $\hat{\mathbf{d}}_n$:

$$(\mathbf{LDL}^H)^{-1} \tilde{\mathbf{z}}_n = \mathbf{d}_n \quad (30)$$

$$\tilde{\mathbf{z}}_n = (\mathbf{LDL}^H) \mathbf{d}_n \quad (31)$$

$$\tilde{\mathbf{z}}_n = \underbrace{\mathbf{L} \mathbf{D} \mathbf{L}^H}_{\mathbf{d}_{2,n}} \mathbf{d}_n \quad (32)$$

The overall complexity for obtaining $\hat{\mathbf{x}}_n$ is $(8Q^2 + 22Q + 4)N_a$ complex operations [42]. The choice of the parameter Q is a trade-off between performance and sophistication. So, for example, a larger Q produces a slight estimation error, resulting in performance enhancement. On the other hand, the calculations' complexity increases as a consequence of the higher bandwidth of \mathbf{B} .

4. Results and discussion

In the following channel environments, the performance of uncoded bit error rate (BER) for the conventional OFDM and DFrFT-OCDM systems is studied:

- 1- Time-invariant channel.
- 2- Time-variant channel.

The QPSK modulated OFDM system under investigation has the following parameters: $L = 8$, $N = 128$, and $N_a = 96$. The communication channel simulated in this proposed work is the Rayleigh fading channel that has an exponential power delay profile and a root-mean-square delay spread of 3. The adopted carrier frequency is chosen to be ultra-high frequency based on the suggested application investigated in this paper, therefore, the subcarrier spacing is $\Delta f = 20$ kHz and $f_c = 10$ GHz. This Doppler frequency corresponds to a high mobile speed $V = 324$ Km/h. Simulation is carried over 10^5 continuous channels and different OFDM symbols, which means $10^5 * 96 * 4$ data bits.

4.1. Time-invariant channel

Doppler frequency is equal to zero in the time invariant channel environment ($f_d = 0$). The OFDM system uses the single tap equaliser, and the DFrFT-OCDM system uses the MMSE equaliser. Fig. 13 shows the BER performance for both systems. The OFDM

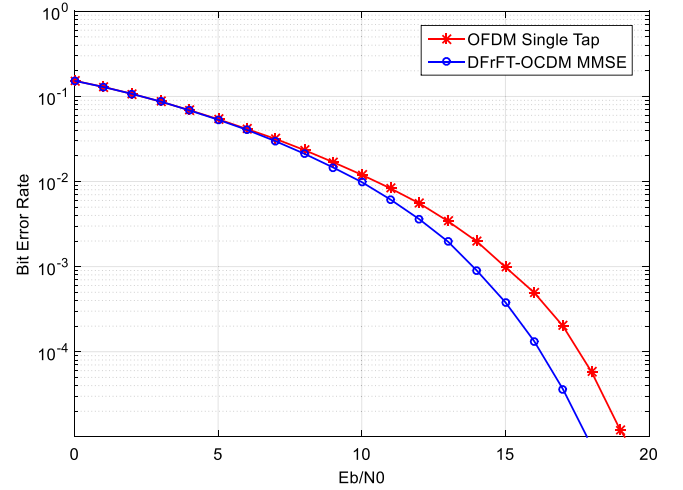


Fig. 12. OFDM and DFrFT-OCDM BER comparison in time invariant channel environment.

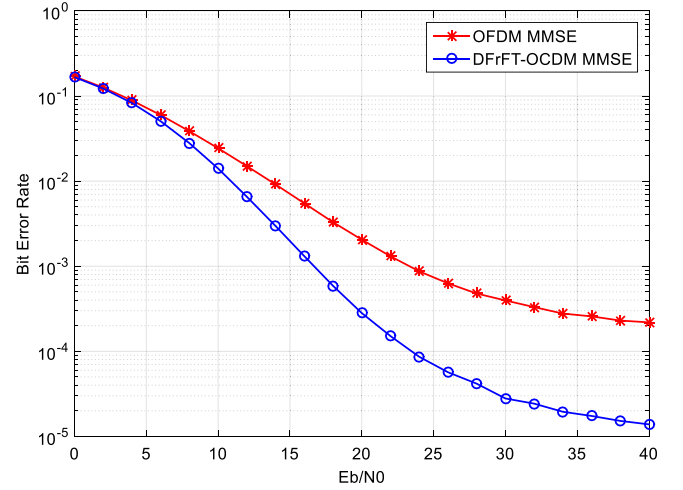


Fig. 13. OFDM and DFrFT-OCDM BER comparison in time-variant channel environment.

system performance is compared to the work carried out in [42], and it was found to match.

From Fig. 12, although the DFrFT-OCDM system has a superior performance, the OFDM system with the single tap equaliser has a very competitive performance with much less complexity. As a result, there is a recommendation to use the OFDM in the time-invariant fading channel scenarios.

4.2. Time-variant channel

In the time-variant channel environment, we consider the maximum Doppler frequency $f_d = 0.15\Delta f$. The MMSE equaliser was used for the OFDM and the DFrFT-OCDM system.

From Fig. 13, the DFrFT-OCDM system has a superior performance when compared to the OFDM system with the same MMSE equaliser, and with the same complexity. As a result, the DFrFT-OCDM is regarded as a better choice in time-variant fading channel scenarios.

4.3. LDL^H factorisation equaliser simulation

To measure the performance of the proposed OFDM system, simulation was run for an OFDM transmission with $N = 128$

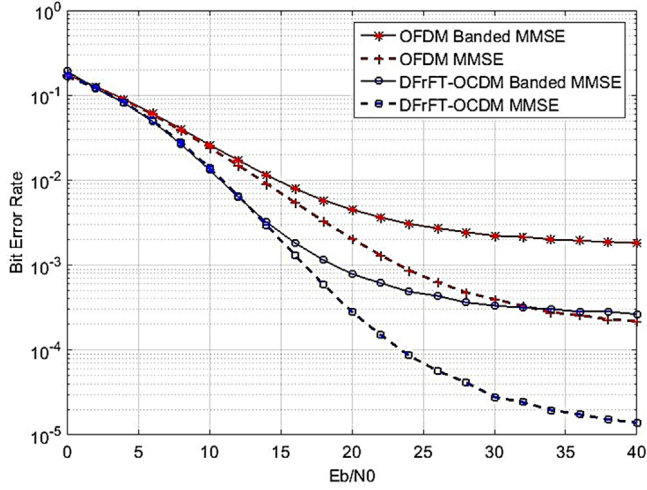


Fig. 14. BER for MMSE equalisation with a block of $Q = N_a(96)$, and low-complexity at $Q = 5$ approaches for DFrFT-OCDM at $\alpha = 0.2\pi/2$ and OFDM.

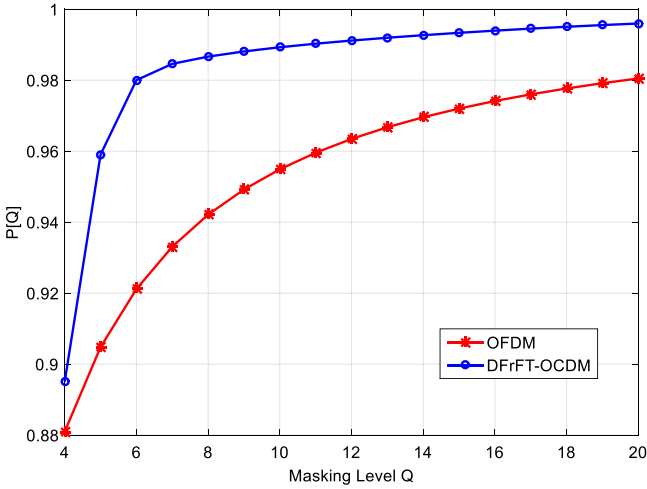


Fig. 15. Comparing OFDM and DFrFT-OCDM at $\alpha = 0.2\pi/2$ for different percentage values of $\tilde{\mathbf{H}}_\alpha$ power restricted in \mathbf{B}_n , determined by $[Q]$ that depends on the number of off-diagonal elements Q that are measured by \mathbf{M} .

subcarriers, where only $N_a = 96$ are active subcarriers, with CP of length $L = 8$, and QPSK modulation. The channel model is the same as the one used in [8], which adapts an exponential power delay profile with an RMS delay spread of 3 sampling periods, with a maximum Doppler spread f_d equal to 15% of the carrier spacing over a group of 10^5 Rayleigh fading channels.

A comparison between the block MMSE equaliser and the LDL^H low-complexity equaliser (Banded MMSE) is shown in Fig. 14 for the OFDM system and the DFrFT-OCDM system ($\alpha = 0.2\pi/2$). The low-complexity equaliser functions with $Q = [5, 96]$, where 96 corresponds to the regular block MMSE equaliser. Performance results are shown in Fig. 14 in terms of BER. The OFDM curves correspond with those stated in [42]. The DFrFT-OCDM system with $Q = 5$ shows a slight degradation over the full MMSE OFDM system at high SNR, but needs only 3.4% [42] of the calculation's cost in terms of complex operations, and still outperforms OFDM.

To investigate and determine the masking level Q 's influence, the power components in $\tilde{\mathbf{H}}_\alpha$ and the after-masking process by \mathbf{M} reduced to \mathbf{B}_n , should be compared. We take into consideration that the ensemble trace operator $tr\{\cdot\}$'s average power ratio is

given in the following relation:

$$\rho[Q] = \varepsilon \left\{ \frac{tr\{\mathbf{B}_n \mathbf{B}_n^H\}}{tr\{\tilde{\mathbf{H}}_\alpha \tilde{\mathbf{H}}_\alpha^H\}} \right\}, \quad 0 \leq \rho[Q] \leq 1 \quad (33)$$

Fig. 15 clearly illustrates that OFDM suffers from the effect of Doppler fading that causes energy to spread away from the main diagonal. The results show that the spread of energy is not even limited to nearby off-diagonals, which makes it essential to imply a high value of Q to collect a large amount of the power contained in $\tilde{\mathbf{H}}$. A similar effect of Doppler fading can be noted with the DFrFT-OCDM because of its inability to diagonalise $\tilde{\mathbf{H}}_\alpha$. However, unlike OFDM, DFrFT-OCDM's leaked power exists adjacent to the off-diagonal element. As a result, a much lower value of Q can be applied to collect the required power of the components of $\tilde{\mathbf{H}}_\alpha$ in \mathbf{B}_n , which justifies the performance improvement achieved with even less difficulty, making the proposal of this system for mobile medical application with machine learning perspectives preferable.

4.4. Deep computing perspectives

Although various concepts of machine learning and big data applications are explored in mobile health, there has been little attention paid to the usage of machine learning and optimisation techniques for coding at the physical layer. At present, the machine learning and deep learning fields have overcome issues concerning compression of neural networks and neural auto-encoders. Consequently, there is adequate opportunity for applying such techniques in the mobile health field; indeed, it could be used in encoding the data prior to broadcasting and then decoding it after passing through the channel. This opportunity is regarded as essential in the sense that a robust and compressed neural net is needed. At the same time, such a model must be able to decrease the load of transmitted data through its auto-encoding structure. Other opportunities consist of the application's design using intelligent models that are specialised for specific medical tasks. They can also communicate through the channel both efficiently and in a compressed manner.

There are various methods that can apply machine learning models for the specific tasks of medical images or video processing [43]. Such models can be quickly and easily learned through different optimisation techniques, such as greedy growing and pruning for trees. These can be easily used for fast auto-encoder tasks because they are compact and simple to interpret. Other types of models, such as neural nets, can be learned through some gradient-based methods. However, for neural nets, due to the high complexity of the model, the issue of compressing and decreasing inference complexity must be tackled for specific tasks of data transmitting. As an extension to the current work, one can further design a model that is suitable for the previously mentioned tasks whilst achieving optimal efficiency in data transmission, encoding, and bandwidth usage.

5. Conclusions

In this paper, a mobile medical video streaming broadcasting system was proposed. The time and frequency fading channel with its effects on OFDM system performance were investigated. The DFrFT-OCDM MCM system was studied as an alternative MCM system that can enhance the overall MCM system performance. It was demonstrated that using simple equalisers with MCM systems was in high demand within the medical video broadcasting system because of its large symbol length, despite its simplicity. The DFrFT-OCDM was found to be a good alternative for the OFDM in a doubly dispersive channel environment,

dependent upon changing the traditional OFDM basis with a chirp basis using the DFrFT, which can cope with the channel variations.

Low-complexity equalisers based on LDL^H factorisation were proposed with the DFrFT-OCDFM system, and it was demonstrated that this new combination shows improved performance when compared to OFDM using the same equalisers. This justifies the reason this system is recommended to be applied within mobile medical video broadcasting. Future work will be implemented to incorporate this proposed system with other techniques that could reduce the complexity or the power consumption, such as searching for new low-complexity equalisers, searching for alternative bases that can improve the MCM system's performance under doubly dispersive fading channels in other applications like social media, introducing Network Coding in [15,44–46] with MCM systems to improve the overall system performance, or with searching for the optimum number of paths for realisation of multi-path routing as in [47].

CRedit authorship contribution statement

Hani H. Attar: Conceptualization, Methodology, Writing - original draft. **Ahmad A.A. Solyman:** Conceptualization, Methodology, Writing - original draft, Investigation, Supervision. **Abd-Elnaser Fawzy Mohamed:** Investigation, Visualization. **Mohammad R. Khosravi:** Resources, Supervision, Writing - review & editing. **Varun G. Menon:** Software, Resources, Writing - review & editing. **Ali Kashif Bashir:** Software, Resources, Writing - review & editing. **Pooya Tavallali:** Resources, Writing - review & editing.

Declaration of competing interest

The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

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