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MMSE channel estimation for two-port demodulation reference signals in new radio

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Dear editor,

Orthogonal frequency division multiplexing (OFDM) has been adopted in new radio (NR) [1], and multiple-port demodulation reference signal (DMRS) has been designed to achieve channel estimation. In NR, two ports of DMRS share the same frequency resource elements (RE) by employing an orthogonal code comprising cyclic shift phase. In general, channel estimation can be achieved via the classical discrete Fourier transform (DFT) scheme [2, 3] as channels for the two ports of DMRS can be separated in the time domain via the DFT scheme. In [4], it has been shown that the minimum mean square error (MMSE) estimator is robust to correlation matrix mismatch; this makes the MMSE estimator more practical. In general, little is known about utilizing the MMSE metric for channel estimation of two-port DMRS in NR.

In this study, we propose a new approach for the channel estimation of two-port DMRS in NR on the basis of the MMSE metric. First, we propose an MMSE with full priori knowledge (F-MMSE) scheme to achieve the channel estimation of two-port DMRS in NR. Then, we present an MMSE with partial priori knowledge (P-MMSE) scheme when the two ports are assigned to different users as full priori knowledge of two ports is not easy to obtain for one user.

Problem formulation. Consider an OFDM system with M subcarriers and equipped with two transmitting and two receiving antennas. Two ports of DMRS share the same time frequency resources. Assume that $2P < M$ subcarriers are assigned to one user, and $x_p = a_{2p,q}$ is the DMRS pilot with transmitting power of 1, where $p = 0, 1, \ldots, P-1$ is the frequency index and q is the time index. $a_{2p+1,q}$ is usually set to zero unless it is used to support more DMRS ports. Hence, in this study, $a_{2p+1,q}$ is set to zero. If x_p is the pilot symbol of the first DMRS port, the pilot symbol of the second DMRS port would be $x_p \mathrm{e}^{\mathrm{j}2\pi\frac{p\Delta_{\mathrm{cs}}}{P}}$ [1], where Δ_{cs} is the cyclic shift and is equal to $\frac{P}{2}$ for the orthogonality of the two DMRS ports.

Then, the demodulation of the rth receiving antenna can be written as follows:

$$y_p^r = h_p^{1r} x_p + h_p^{2r} x_p \mathrm{e}^{\mathrm{j}2\pi\frac{p\Delta_{\mathrm{cs}}}{P}} + \eta_p^r, \tag{1}$$

where $r = 0$ or 1, h_p^{tr} is the channel frequency response for the pth pilot between the tth transmitting antenna and rth receiving antenna, and η_p^r is the additive white Gaussian noise with mean zero and variance σ^2.

For simplicity, the matrix form of (1) can be expressed as

$$\boldsymbol{y}_r = \boldsymbol{X}\boldsymbol{h}_{1r} + \boldsymbol{X}\boldsymbol{C}\boldsymbol{h}_{2r} + \boldsymbol{\eta}_r, \tag{2}$$

where $\boldsymbol{y}_r = [y_0^r, y_1^r, \ldots, y_{P-1}^r]^{\mathrm{T}}$, $\boldsymbol{h}_{tr} = [h_0^{tr}, h_1^{tr}, \ldots, h_{P-1}^{tr}]^{\mathrm{T}}$, $\boldsymbol{\eta}_r = [\eta_0^r, \eta_1^r, \ldots, \eta_{P-1}^r]^{\mathrm{T}}$, and $\boldsymbol{X} = \mathrm{diag}(x_0, \ldots, x_p, \ldots, x_{P-1})$. $\boldsymbol{C}$ is a diagonal matrix and its the pth diagonal element is $\mathrm{e}^{\mathrm{j}2\pi\frac{p\Delta_{\mathrm{cs}}}{P}}$.

Then, the DMRS-based channel estimation model can be expressed as follows:

$$\hat{\boldsymbol{h}} = \boldsymbol{X}^{-1}\boldsymbol{y}_r = \boldsymbol{h}_{1r} + \boldsymbol{C}\boldsymbol{h}_{2r} + \boldsymbol{X}^{-1}\boldsymbol{\eta}_r, \tag{3}$$

where $\boldsymbol{X}^{-1}$ is the inverse matrix of $\boldsymbol{X}$.

Our algorithm. The F-MMSE channel estimation of $\boldsymbol{h}_{1r}$ can be obtained by minimizing the mean square error (MSE), which is defined as

$$\mathrm{MSE} = \mathbb{E}\left\{\left\|\boldsymbol{h}_{1r} - \hat{\boldsymbol{h}}_{1r}\right\|_2^2\right\}, \tag{4}$$

where $\hat{\boldsymbol{h}}_{1r}$ is an estimate of $\boldsymbol{h}_{1r}$. Subsequently, the estimation of $\boldsymbol{h}_{1r}$ by F-MMSE is obtained by minimizing the MSE:

$$\hat{\boldsymbol{h}}_{1r}^{\text{F-MMSE}} = \boldsymbol{R}_{\hat{\boldsymbol{h}}\boldsymbol{h}_{1r}} \boldsymbol{R}_{\hat{\boldsymbol{h}}\hat{\boldsymbol{h}}}^{-1} \hat{\boldsymbol{h}}, \tag{5}$$

where $\boldsymbol{R}_{\hat{\boldsymbol{h}}\boldsymbol{h}_{1r}}$ is the cross covariance matrix between $\hat{\boldsymbol{h}}$ and $\boldsymbol{h}_{1r}$. Furthermore, $\boldsymbol{R}_{\hat{\boldsymbol{h}}\hat{\boldsymbol{h}}}$ is the autocovariance matrix of $\hat{\boldsymbol{h}}$. $\mathbf{R}_{\hat{\boldsymbol{h}}\boldsymbol{h}_{1r}}$ and $\boldsymbol{R}_{\hat{\mathbf{h}}\hat{\mathbf{h}}}^{-1}$ are obtained as

$$\boldsymbol{R}_{\hat{\boldsymbol{h}}\boldsymbol{h}_{1r}} = \mathbb{E}\left\{\boldsymbol{h}_{1r}\hat{\boldsymbol{h}}^{\dagger}\right\} = \boldsymbol{R}_{1r}, \tag{6}$$

$$\boldsymbol{R}_{\hat{\boldsymbol{h}}\hat{\boldsymbol{h}}} = \mathbb{E}\left\{\hat{\boldsymbol{h}}\hat{\boldsymbol{h}}^{\dagger}\right\} = \boldsymbol{R}_{1r} + \boldsymbol{C}\boldsymbol{R}_{2r}\boldsymbol{C}^{\dagger} + \sigma^2\boldsymbol{I}_P, \tag{7}$$

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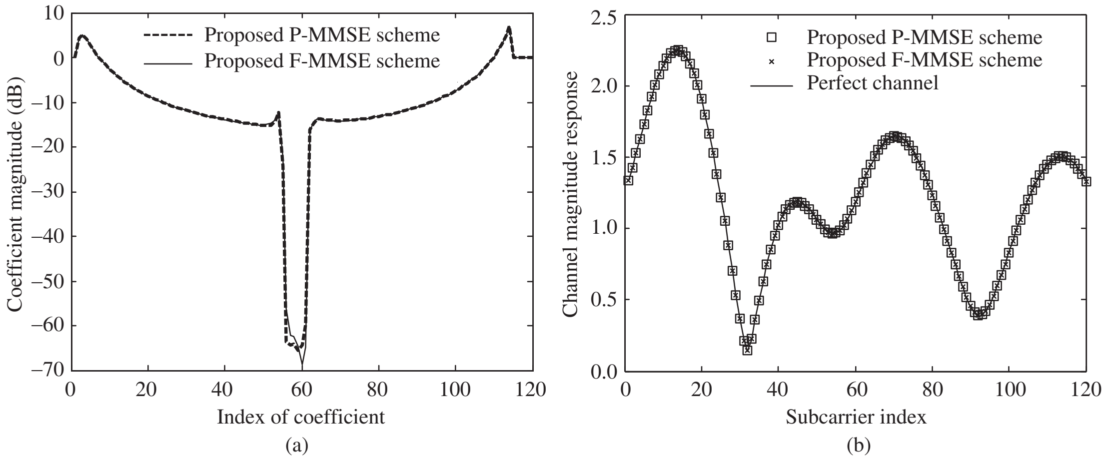

Figure 1 Comparison of P-MMSE and F-MMSE, SNR = 30 dB. (a) Diagonal elements of coefficient matrices; (b) channel magnitude responses.

where $\boldsymbol{I}_P$ is the identity matrix with dimension of P and $(\cdot)^\dagger$ is the operation of the conjugate transpose. $\boldsymbol{R}_{1r}$ and $\boldsymbol{R}_{2r}$ are autocovariances of $\boldsymbol{h}_{1r}$ and $\boldsymbol{h}_{2r}$. Then, the F-MMSE can be obtained as

$$\hat{\boldsymbol{h}}_{1r}^{\text{F-MMSE}} = \boldsymbol{R}_{1r}\left(\boldsymbol{R}_{1r} + \boldsymbol{C}\boldsymbol{R}_{2r}\boldsymbol{C}^\dagger + \sigma^2\boldsymbol{I}_P\right)^{-1}\hat{\boldsymbol{h}}. \tag{8}$$

Although the abovementioned F-MMSE is straightforward, it helps us to develop a more practical estimator. When the two DMRS ports are assigned to different users, one user performing the channel estimation does not have the priori knowledge of the other DMRS port or that whether the other DMRS port is being used. In the proposed P-MMSE scheme, we replace $\boldsymbol{R}_{2r}$ with $\boldsymbol{R}_{1r}$ and the channel estimation is written as

$$\hat{\boldsymbol{h}}_{1r}^{\text{P-MMSE}} = \boldsymbol{R}_{1r}\left(\boldsymbol{R}_{1r} + \boldsymbol{C}\boldsymbol{R}_{1r}\boldsymbol{C}^\dagger + \sigma^2\boldsymbol{I}_P\right)^{-1}\hat{\boldsymbol{h}}. \tag{9}$$

Performance analysis. To demonstrate the validity of the proposed P-MMSE scheme, a theoretical analysis is presented. First, the frequency-domain estimation in (9) is converted to time domain as

$$\begin{aligned}\hat{\underline{\boldsymbol{h}}}_{1r}^{\text{P-MMSE}} &= \boldsymbol{F}\hat{\boldsymbol{h}}_{1r}^{\text{P-MMSE}} \\ &= \boldsymbol{\Phi}_{1r}^{\text{P-MMSE}}\left(\underline{\boldsymbol{h}}_{1r} + \underline{\boldsymbol{h}}_{2r}\left(\frac{P}{2}\right)\right) + \boldsymbol{\Phi}_{1r}^{\text{P-MMSE}}\boldsymbol{\xi}_r,\end{aligned} \tag{10}$$

where $\boldsymbol{\Phi}_{1r}^{\text{P-MMSE}} = \boldsymbol{F}\boldsymbol{R}_{1r}\left(\boldsymbol{R}_{1r} + \boldsymbol{C}\boldsymbol{R}_{1r}\boldsymbol{C}^\dagger + \sigma^2\boldsymbol{I}_P\right)^{-1}\boldsymbol{F}^\dagger$. Note that the coefficient matrix of F-MMSE is $\boldsymbol{\Phi}_{1r}^{\text{F-MMSE}} = \boldsymbol{F}\boldsymbol{R}_{1r}\left(\boldsymbol{R}_{1r} + \boldsymbol{C}\boldsymbol{R}_{2r}\boldsymbol{C}^\dagger + \sigma^2\boldsymbol{I}_P\right)^{-1}\boldsymbol{F}^\dagger$. $\boldsymbol{F}$ is the IDFT matrix with dimension of $P \times P$ whose the (p,q)th element is $\frac{1}{\sqrt{P}}e^{j2\pi\frac{pq}{P}}$, and $\boldsymbol{\xi}_r = \boldsymbol{F}\boldsymbol{X}^{-1}\boldsymbol{\eta}_r$. $\underline{\boldsymbol{h}}_{tr}$ is obtained by performing IDFT on $\boldsymbol{h}_{tr}$, and $\underline{\boldsymbol{h}}_{2r}(\frac{P}{2})$ is a cyclic shift vector of $\underline{\boldsymbol{h}}_{2r}$ with a shift $\frac{P}{2}$.

It can be observed from (10) that the estimation $\hat{\underline{\boldsymbol{h}}}_{1r}^{\text{P-MMSE}}$ is a linear combination of $\underline{\boldsymbol{h}}_{1r}$ and $\underline{\boldsymbol{h}}_{2r}(\frac{P}{2})$ by the P-MMSE coefficient $\boldsymbol{\Phi}_{1r}^{\text{P-MMSE}}$. It can be seen that the values of the channel estimation depend mainly on the diagonal coefficients of the matrix $\boldsymbol{\Phi}_{1r}^{\text{P-MMSE}}$. Figure 1(a) depicts the magnitudes of the diagonal elements of the P-MMSE and F-MMSE coefficient matrices, where $P = 120$ and $M = 2048$. In addition, a random channel model is adopted with 40 sample-spaced independent Rayleigh fading paths, which exhibits an exponential power delay profile [5] as $\alpha(l) = e^{\beta l}$, where $l = 0, 1, \ldots, 39$, $\beta = -0.0005$ for the first DMRS port and $\beta = -0.05$ for the second DMRS port, i.e., $\beta = -0.0005$ is for the channels $\boldsymbol{h}_{11}$ and $\boldsymbol{h}_{12}$, and $\beta = -0.05$ is for the channels $\boldsymbol{h}_{21}$ and $\boldsymbol{h}_{22}$. Different β in channels lead to different channel covariance matrices. From the simulation results, it is evident that the magnitude of a diagonal element of the P-MMSE coefficient matrix is in line with the F-MMSE despite a large difference in the attenuation factor β, which is insensitive to channel covariance matrices. Furthermore, it can be observed that the P-MMSE coefficients have very small values at the points around $\frac{P}{2}$, which indicates that for the channel estimation of $\underline{\boldsymbol{h}}_{1r}$, the interference from $\underline{\boldsymbol{h}}_{2r}(\frac{P}{2})$ is significantly suppressed as the channel coefficients of $\underline{\boldsymbol{h}}_{2r}(\frac{P}{2})$ have valid values at the points around $\frac{P}{2}$. From Figure 1(b), it can be observed that the channel estimations of the proposed P-MMSE and F-MMSE schemes are in line with the perfect channel $\boldsymbol{h}_{11}$, without significant interference.

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Supporting information Appendixes A–C. The supporting information is available online at info.scichina.com and link.springer.com. The supporting materials are published as submitted, without typesetting or editing. The responsibility for scientific accuracy and content remains entirely with the authors.

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