ESPAR Antenna Based Diversity Scheme for Multiple Antenna OFDM System

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Abstract—**This paper proposes a diversity scheme for multiple antenna in OFDM system based on an electrically steerable parasitic array radiator (ESPAR) antenna. Although the 2 x 2 OFDM system is capable of doubling the capacity without expanding the occupied frequency bandwidth, we can't get the additional diversity gain using the linear MIMO decomposition method. This proposed method can improve the bit error rate performance (BER) by making efficient use of ESPAR antenna. Computer simulation results shows that the proposed scheme gives the additional diversity gain about 3.9 dB for ideal channel and 2.55 dB for estimated channel for maximal likelihood detection (MLD) with scheme 2 x 2 MIMO conventional.**

Keywords—OFDM, ESPAR, MIMO-decomposition, bit error rate.

I. INTRODUCTION

MIMO (Multiple Input Multiple Output) term or we say multiple antenna is a technique to increase the transmission bit rate without extending frequency bandwidth [1],[4],[8] and [9]. However, multiple antennas require the same number of Radio Frequency (RF) front-end circuits. A single RF MIMO system based on ESPAR (Electrically Steerable Parasitic Array Radiator) antenna has been proposed in [5], transmitter for multiple antennas in wireless communication using a single RF front-end at the mobile terminal. Although in [5] can reduce the number of RF front-end, it couldn't apply to multiple antennas for OFDM system. Furthermore, the performance in [5] for system 2 x 2 using 3 and 5 element ESPAR antenna still worse than multiple antenna in conventional system. A part of authors has proposed a new diversity scheme based on ESPAR antenna [7], the schemes based on ESPAR antenna, whose beam direction is oscillated in the symbol time of the received OFDM signal. This schemes gives diversity gain in a frequency selective fading channels, but the system present in [7] is still single antenna system.

In this paper, we propose a multiple antenna for OFDM system with ESPAR antenna at receiver side in order to improve the bit error rate performance without sacrificing the transmission data rate. Computer simulation results that the proposed scheme is capable of obtaining additional diversity gain in frequency selective fading without reducing the special multiplexing capability. But the complexity of computational cost of MIMO decomposition is still high.

II. ESPAR ANTENNA BASED DIVERSITY RECEIVER FOR OFDM

The proposed multiple antennas for OFDM system with ESPAR antenna are an extension of the ESPAR antenna-based diversity receiver proposed by a part of authors in [7]. In this section, we will briefly introduce the ESPAR antenna-based diversity receiver for OFDM. Fig. 1 illustrates the block diagram of transmitter and receiver system of OFDM ESPAR system. ESPAR antenna in this section is composed of a radiator and parasitic elements terminated with a variable capacitor. Since the radiator and parasitic elements are electro-magnetically coupled, the output of the ESPAR antenna is a weighted sum of the received signal at the each element.

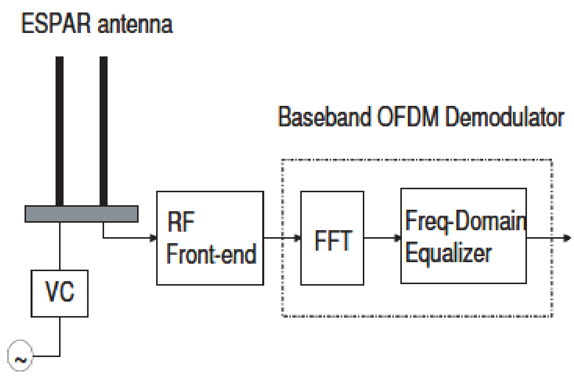

Fig.1. ESPAR antenna-based diversity receiver for OFDM system

Assume $v_d(t)$ and $v_p(t)$ be the open circuit voltage of the radiator and parasitic elements, respectively. The output signal is shown

$$v(t) = \alpha v_d(t) + \beta v_p(t) \tag{1}$$

where α and β are the weighting factors determine by the

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reactance value of the variable capacitor terminating the parasitic elements. The principle ESPAR antenna is introduced in [6]. In this proposal the reactance varies by the oscillator whose frequency is the same as the subcarrier spacing of the OFDM signal. Then the output of the ESPAR antenna can be rewritten as

$$v(t) = \alpha v_d(t) + \beta e^{j2\pi t/T_s} v_p(t) \qquad (2)$$

Where T_s is a FFT (fast Fourier transform) window period of the OFDM signal. At the receiver, special equalization is required for reducing the Inter Channel Interference (ICI) generated by proposed scheme. In the following we derive the channel estimation algorithm for the proposed ESPAR antenna based diversity receiver. The received signal is applied to the FFT processor followed by the frequency domain equalizer. Now let assume that the transmitted symbol vector in the frequency domain is

$$\mathbf{x} = [x_0, x_1, x_2, \cdots, x_{N-1}]^T \qquad (3)$$

where x_k is data symbols of kth subcarriers, N is a number of subcarriers. The data symbol in time domain is given by

$$\mathbf{v} = \mathbf{F}^{-1}\mathbf{x} \qquad (4)$$

where **F** is the Fourier transformation matrix. That is, the kth column lth raw element of **F** is $e^{j2\pi kl/N}$. The transmitted symbol **v** is then propagated through the multipath fading channel. The received signal component at the ith element is given by

$$\mathbf{q}_i = \mathbf{C}_i \mathbf{v} \qquad (5)$$

Where $\mathbf{C}_i$ is the channel impulse response matrix corresponding the ith element. The received signal at the input of the baseband demodulation block is given by

$$\begin{aligned} \mathbf{r} &= \mathbf{q}_o + \mathbf{D}\mathbf{q}_i + \mathbf{z} \\ &= (\mathbf{C}_o + \mathbf{D}\mathbf{C}_1)\mathbf{v} + \mathbf{z} \\ &= (\mathbf{C}_o + \mathbf{D}\mathbf{C}_1)\mathbf{F}^{-1}\mathbf{x} + \mathbf{z} \end{aligned} \qquad (6)$$

where **z** is the thermal noise component, and **D** is diagonal d_k is a diagonal matrix whose kth diagonal element is d_k. According to (2), d_k is given by

$$d_k = e^{2\pi k/N} \qquad (7)$$

The received signal is applied to FFT. The output of FFT is then given by

$$\mathbf{u} = \mathbf{Fr}(\mathbf{H}_o + \mathbf{G}\mathbf{H}_1)\mathbf{x} + \mathbf{z} \qquad (8)$$

where

$$\mathbf{H}_i = \mathbf{F}\mathbf{C}_i\mathbf{F}^{-1} \qquad (9)$$

is a diagonal matrix, whose diagonal elements represents the frequency response.

$$\mathbf{G} = \mathbf{F}\mathbf{D}\mathbf{F}^{-1} \qquad (10)$$

where **G** represents Inter-Channel Interference (ICI) matrix, and **z** is the thermal noise component in the frequency domain.

A. Channel Estimation

According to the (3), the received signal in the proposed scheme was interfered by the adjacent subcarrier signals. That is, the frequency domain equalization is required at the receiver. It is required to estimate the frequency response of the channel for equalizing the signal. In order to estimate the frequency response, the proposed scheme first transmits a pilot symbol before transmitting the data. Now let us assume that the pilot symbol vector in the frequency domain is

$$\mathbf{p} = [p_o, p_1, p_2, \cdots, p_{N-1}]^T \qquad (11)$$

Where p_k is data symbols of kth subcarriers, N is a number of subcarriers. The data symbol in time domain is given by

$$\mathbf{u} = \mathbf{Fr}(\mathbf{H}_o + \mathbf{G}\mathbf{H}_1)\mathbf{p} + \mathbf{z} \qquad (12)$$

Now, let us assume that the vector **p** and $\mathbf{h}_i$ represent the diagonal components of the matrices **P** and $\mathbf{H}_i$, respectively. Now, let us assume that the vector **p** and h_i represent the diagonal components of the matrices **P** and $\mathbf{H}_i$, respectively. That is,

$$\mathbf{P} = \mathrm{diag}(\mathbf{p}) \qquad (13)$$

and

$$\mathbf{H}_i = \mathrm{diag}(\mathbf{h}_i) \qquad (14)$$

Then without loss of generality, we can exchange the diagonal matrices and vectors as in

$$\mathbf{u} = \mathbf{P}\mathbf{h}_o + \mathbf{G}\mathbf{P}\mathbf{h}_1 + \mathbf{z} \qquad (15)$$

The covariance and cross-correlation matrices are now derived as

$$\begin{aligned} \mathbf{R} &= E[\mathbf{u}\mathbf{u}^H] \\ &= \mathbf{P}\mathbf{R}_h\mathbf{P}^H + \mathbf{G}\mathbf{P}\mathbf{R}_h\mathbf{P}^H\mathbf{G}^H + \sigma_z^2\mathbf{I} \end{aligned} \qquad (16)$$

And

$$\mathbf{B}_i = E[\mathbf{u}\mathbf{h}_i^H] \qquad (17)$$

where

$$\mathbf{R}_h = E[\mathbf{h}_i\mathbf{h}_i^H] \qquad (18)$$

is the covariance matrix of the channel response and $E[.]$ denotes the ensemble average. The cross-correlation matrices are further given by

$$\mathbf{B}_o = \mathbf{PR}_h \tag{19}$$

and

$$\mathbf{B}_1 = \mathbf{GPR}_h \tag{20}$$

The channel response is finally estimated by

$$\mathbf{h}_i = \mathbf{W}_i^H \mathbf{u} \tag{21}$$

where $\mathbf{W}_i = \mathbf{R}^+ \mathbf{B}_i$ is the weight matrix in terms of minimum mean square error (MMSE) criterion, where R^+ is the pseudo inverse matrix of **R**.

B. Frequency Domain Equalizer

From the estimated channel response, we can equalize the received data symbols. In the following we assume the simple zero forcing (ZF) equalizer. The estimated channel response is given by

$$\mathbf{x} = \widehat{\mathbf{H}}_o + \mathbf{G}\widehat{\mathbf{H}}_1 \tag{22}$$

where $\widehat{H}_l$ is estimated channel response.

III. ESPAR ANTENNA BASED MULTIPLE ANTENNA IN OFDM SYSTEM

In this section we present multiple antena in OFDM system for size of 2 x 2 with ESPAR antenna in the receiver side. The transmitter and receiver block diagrams of the proposed shown in Fig. 2 and Fig. 3. Fig. 2 represents the pilot symbol transmission phase of transmitter. As introduced in the conventional MIMO-OFDM WLAN standar, IEEE 802.11n, the pilot tone for the seconds stream of the transmitter is shifted by CS (Cyclic Shift) block [2].

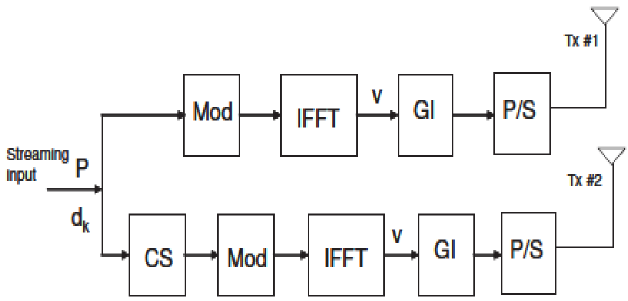

Fig.2. Block diagram of transmitter multiple antenna in OFDM system

At the receiver in Fig.3, we use two ESPAR antena for receiving the signal. The received signals are then applied to the corresponding FFT processor followed by the channel estimator and MIMO decoder. The output of the MIMO decoder is then applied to the demapper to demodulate the symbol.

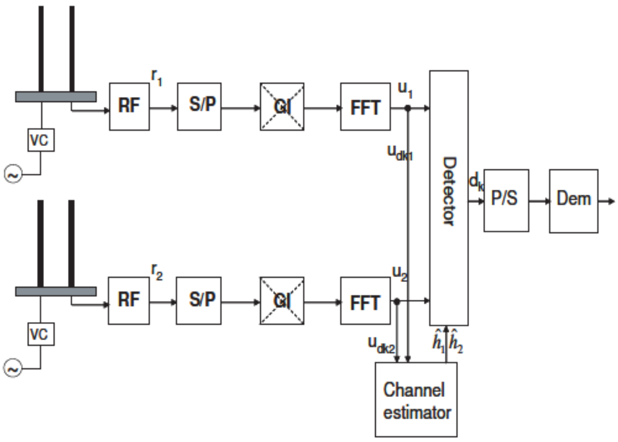

Fig.2. Block diagram of receiver for multiple antenna in OFDM system using ESPAR antenna

A. Channel Estimator

Let the pilot symbol and its cyclic shifted one be $\mathbf{P}_1$ and $\mathbf{P}_2$, respectively. Then the received signal $\mathbf{u}_i$ are calculated as in [7] at the *i*-th ESPAR antenna in the frequency domain is given by

$$\mathbf{u}_i = \mathbf{P}_1\mathbf{h}_{2i,0} + \mathbf{GP}_1\mathbf{h}_{2i+1,0} + \mathbf{P}_2\mathbf{h}_{2i,1} + \mathbf{GP}_2\mathbf{h}_{2i+1,0} + \mathbf{z} \tag{23}$$

By using $\mathbf{R} = E[\mathbf{u}_i \mathbf{u}_i^H]$ and $\mathbf{B}_i = E[\mathbf{u}_i \mathbf{h}_i^H]$ for auto-correlation and cross-correlation, respectively, then from (23) we can obtain the auto-correlation matrix formed by

$$\mathbf{R} = \mathbf{P}_1\mathbf{R}_h\mathbf{P}_1^H + \mathbf{GP}_1\mathbf{R}_h\mathbf{P}_1^H\mathbf{G}^H + \mathbf{P}_2\mathbf{R}_h\mathbf{P}_2^H + \mathbf{GP}_2\mathbf{R}_h\mathbf{P}_2^H\mathbf{G}^H + \sigma_Z^2\mathbf{I} \tag{24}$$

And cross-correlation matrix formed by

$$\left.\begin{array}{lcl} \mathbf{B}_{10} & = & \mathbf{P}_1\mathbf{R}_h \\ \mathbf{B}_{11} & = & \mathbf{GP}_1\mathbf{R}_h \\ \mathbf{B}_{20} & = & \mathbf{P}_2\mathbf{R}_h \\ \mathbf{B}_{21} & = & \mathbf{GP}_2\mathbf{R}_h \end{array}\right\} \tag{25}$$

where $\mathbf{B}_{10}$, $\mathbf{B}_{11}$, and $\mathbf{B}_{20}$, $\mathbf{B}_{21}$ are cross correlation matrix for first pilot and second pilot, respectively and $\mathbf{R}_h$ is correlation inter channel. In our research, we use a rectangular shaping multipath power delay profile as shown in (26) in frequency domain, where Δf denotes the subcarrier spacing and τ_{rms} denotes root mean squared delay spread.

$$\mathbf{R}_h(k) = e^{-j\pi\Delta\tau_{rms}\frac{\sin(\pi\Delta\tau_{rms}k)}{\pi\Delta\tau_{rms}k}} \tag{26}$$

The channel response for 2 x 2 correspondences to first receiver and second receiver respectively is finally estimated by

$$\left.\begin{array}{lll}\mathbf{h}_{10} & = & \mathbf{W}_{10}^{H}\mathbf{u}_{1} \\ \mathbf{h}_{11} & = & \mathbf{W}_{11}^{H}\mathbf{u}_{1} \\ \mathbf{h}_{20} & = & \mathbf{W}_{20}^{H}\mathbf{u}_{1} \\ \mathbf{h}_{21} & = & \mathbf{W}_{21}^{H}\mathbf{u}_{1}\end{array}\right\} \quad (27)$$

$$\left.\begin{array}{lll}\mathbf{h}_{30} & = & \mathbf{W}_{10}^{H}\mathbf{u}_{2} \\ \mathbf{h}_{31} & = & \mathbf{W}_{11}^{H}\mathbf{u}_{2} \\ \mathbf{h}_{40} & = & \mathbf{W}_{20}^{H}\mathbf{u}_{2} \\ \mathbf{h}_{41} & = & \mathbf{W}_{21}^{H}\mathbf{u}_{2}\end{array}\right\} \quad (28)$$

where

$$\left.\begin{array}{lll}\mathbf{W}_{10} & = & \mathbf{R}^{+}\mathbf{B}_{10} \\ \mathbf{W}_{11} & = & \mathbf{R}^{+}\mathbf{B}_{11} \\ \mathbf{W}_{20} & = & \mathbf{R}^{+}\mathbf{B}_{20} \\ \mathbf{W}_{21} & = & \mathbf{R}^{+}\mathbf{B}_{21}\end{array}\right\} \quad (29)$$

B. MIMO decoder

After estimation of channel response, MIMO decoding is performed. In the following we use a Vertical Bell Layer Space Time (V-BLAST) based algorithm, which performs the MIMO decoding and frequency domain. In the following, we assume the number of subcarriers used data transmission is *N*. The ESPAR antenna-based diversity receiver in sec. II, the inter-channel interference is generated at the ESPAR antenna. Therefore, the number of received symbol is $N+2$. The block diagram of detection processing V-BLAST processor is shown in Fig. 4. The V-BLAST processor performs MIMO decoding and frequency domain equalization simultaneously. The channel matrix of frequency components be given by (31) where $\mathbf{H}_{ij}$ is channel matrix for receiver *j* from transmitter *i*, and the size of each matrix are $(2+N)$ x N, hence the matrix size of $\mathbf{H}$ are $2(N+2)$ x $2N$. $\mathbf{H}_{ij}$ form is shown in (30).

$$\mathbf{H} = \begin{bmatrix} H_{11} & H_{12} \\ H_{21} & H_{22} \end{bmatrix} \quad (31)$$

The algorithm utilizes the linear nulling and successive interference cancellation to estimate *N* transmitted symbols from the received signal vector [3]. The symbol is first detected using a linear process such as zero-forcing (ZF) or minimum mean squares error (MMSE). The detected symbol is regenerated, and the corresponding signal portion is subtracted from the received signal vector with the fewer interfering signal component. This process is repeated, until all *N* symbols are detected.

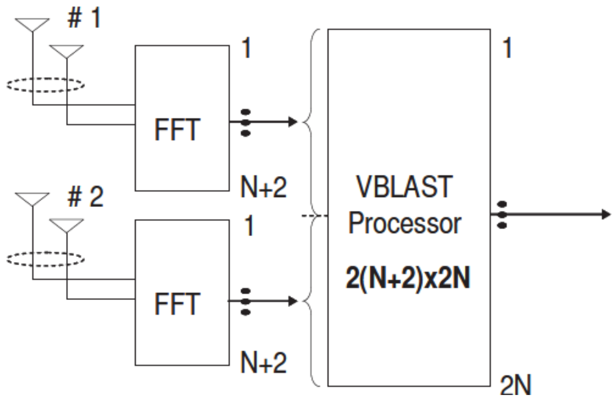

Fig.4. Block diagram of detection using V-BLAST Processor

Let assume $\mathbf{a} = [a_1, a_2, a_3, \cdots, a_{N/2}]^T$ be the transmitted symbol vector, then the received symbol is $\mathbf{r} = \mathbf{Ha} + \mathbf{v}$, where **H** is either one of the channel matrix given in (31). Assume $\mathbf{S} = \{k_1, k_2, k_3, \cdots, k_{N/2}\}$ denotes sequence of sub-stream data received, and then detection process as follows:

Initialization: (32)

$$i \leftarrow 1$$
$$\mathbf{G}_1 = \mathbf{H}^{+}$$
$$k_1 = \underset{\min j}{\arg} \left\| (G_1)_j \right\|^2$$

Recursion: (33)

$$w_{ki} = (\mathbf{G}_i)_{ki}$$
$$\mathbf{y}_{ki} = w_{ki}\mathbf{r}_i$$
$$\mathbf{r}_{i+1} = \mathbf{r}_i - \hat{a}_{ki}\mathbf{H}_{ki}$$
$$\mathbf{G}_{i+1} = \mathbf{H}_{ki}^{+}$$
$$k_{i+1} = \underset{\min j \in \{k1, k2, \cdots kN\}}{\arg} \left\| (\mathbf{G}_{i+1}) \right\|^2$$

IV. Simulations and Discussions

A. Sistem Parameter

The transmitted symbols belong to a QPSK constellation and are obtained from the information bits. Block diagram for MIMO detection in our simulation using Fig. 5. In this simulation we use pilot sequence of High Throughput Long Training Field (HTLTF) based on IEEE.802.11n for frequency is 20 MHz. The total number of subcarriers are 56 and complete of simulation parameter is shown in Table 1. The multipath fading channel is generated by Rayleigh fading channel [9].

Table 1
Parameter System

	Parameter	Value
Transmitter	Pilot sequence	HTLTF for 20 MHz
	Type of Modulation	QPSK
	Number of subcarrier	56
	Size of FFT/IFFT	64
	Antenna dimension	2x2
	Guard interval ratio	¼
Channel	Rayleigh fading	Two-rays
Receiver	Channel estimation	MMSE
	Equalization	V-BLAST

B. Performance Assesment

This section we show output our scheme simulation for a 2 x 2 OFDM ESPAR. The results of the simulation under two-rays Rayleigh fading channel is shown in Fig.6.

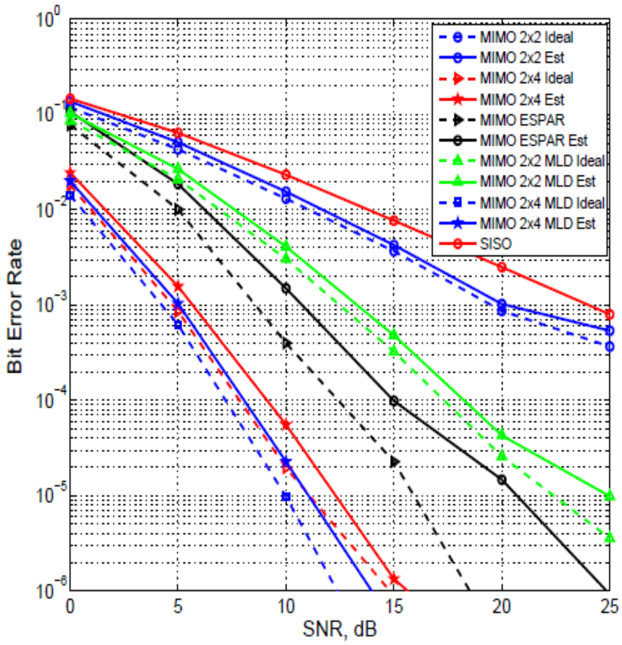

Fig.6. BER performance of multiple antenna OFDM using ESPAR antennas under Rayleigh fading channel

The performances are compared to the corresponding performance of MIMO-OFDM conventional system for 2x4, 2x2 and SISO system, respectively using the same modulation schemes. In Fig. 6 we compare performance bit error rate our scheme for various MIMO detector and our proposal get improvement bit error rate of 10^{-3} gives diversity gain of 10.9 dB for ideal channel and 9.45 dB for estimated channel for V-BLAST (2 x 2). Our proposal so gives diversity gain of 3.9 dB for ideal channel and 2.55 dB for estimated channel for maximal likelihood detection (MLD) with scheme 2 x 2. Our scheme yields results better than a 2 x 2 MIMO conventional system and SISO system. As it is known that a 2 x 2 MIMO conventional system can't get diversity gain as a 2 x 4 MIMO conventional [4], but our scheme can yields better performance for increasing diversity gain than a 2 x 2 MIMO-OFDM conventional. Our scheme is slightly worse than a 2 x 4 MIMO-OFDM conventional.

V. CONCLUSION

This work has shown that multiple antenna for OFDM system using ESPAR antenna in receiver side, it is possible to achieve spectral efficiency gain comparable. Computer simulations results was shown for performance verification of the proposed scheme. It was confirmed that the proposed schemes gives diversity gain in a frequency selective fading channel.

VI. ACKNOWLEDGMENT

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$$
H_{ij} = \begin{pmatrix}
H_{0,-\frac{N}{2}} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
H_{1,-\frac{N}{2}} & H_{0,-\left(\frac{N}{2}-1\right)} & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & H_{1,-\left(\frac{N}{2}-1\right)} & H_{0,-\left(\frac{N}{2}-2\right)} & 0 & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & & H_{1,-\left(\frac{N}{2}-2\right)} & \ddots & 0 & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & \ddots & H_{0,-1} & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & H_{1,-1} & 0 & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & H_{0,+1} & 0 & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & H_{1,+1} & \ddots & 0 & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & \ddots & H_{0,+\left(\frac{N}{2}-2\right)} & 0 & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & H_{1,+\left(\frac{N}{2}-2\right)} & H_{0,+\left(\frac{N}{2}-1\right)} & 0 \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & H_{1,+\left(\frac{N}{2}-2\right)} & H_{0,+\frac{N}{2}} \\
0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & 0 & H_{1,+\frac{N}{2}}
\end{pmatrix}
\tag{30}
$$