A Space-Time Code Design for Omnidirectional Transmission in Massive MIMO Systems

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***Abstract*—Omnidirectional space-time coding has been proposed recently for massive multiple-input multiple-output (MIMO) systems to broadcast common information. In these systems, signals across all transmit antennas have constant power and their discrete Fourier transforms have constant power as well, where the number of discrete frequencies is the number of transmit antennas. When the number of transmit antennas is large, i.e., it is a massive MIMO system, signals received at these discrete angles have constant mean power at any instant time. Furthermore, spatial diversity can be achieved. In this letter, we propose a different design of such a transmission, where the sum of received signal powers in a few consecutive time slots is constant at any angle (not just discrete angles). The other two properties mentioned above, namely, constant transmission power across antennas at any instant time and achieving spatial diversity, are still maintained. Furthermore, it has the fast symbol-wise maximum-likelihood decoding.**

***Index Terms*—Massive MIMO, space-time block code (STBC), common information, omnidirectional transmission, polyphase complementary sequences, lossless matrices.**

I. Introduction

Massive multiple-input multiple-output (MIMO) systems have been commonly recognized as potential systems in the 5th generation (5G) cellular applications and have attracted much attention lately [1], [2]. For broadcasting common information from a base station (BS) in a massive MIMO system, omnidirectional space-time coding has been recently proposed in [4]–[7]. In these systems, all the signal powers across all the transmission antennas are constant and spatial diversity can be achieved. In [4], received mean signal power at finite discrete angles is constant. In [5]–[7], received signal power at any instant time is constant at finite discrete angles. The number of the finite discrete angles is the same as the number of transmit antennas.

In this letter, we propose a different design of omnidirectional space-time coding, where the other two properties of the design are the same as those in [5]–[7], namely the constant transmission power across all the transmit antennas at any instant time, and achieving spatial diversity. At the receiver side, however, the sum of received signal powers at a few consecutive time slots is constant at any angle (not just at finite many discrete angles).

The remainder of this letter is organized as follows. In Section II, the problem of interest is presented, where three design criteria are formulated. In Section III, a concrete STBC design is proposed, which satisfies all the three criteria. In Section IV, lossless polyphase matrices are introduced with their factorizations that will be potentially useful for the STBC design problem in this letter. In Section V, this letter is concluded. In what follows, boldface capital letters stand for matrices and boldface lowercase letters stand for vectors.

II. Problem Description

In this letter, we adopt all the notations and signal models from [5]–[7]. Consider a BS with M transmit antennas and M is large. Let $\mathbf{h} \in \mathbb{C}^{M\times 1}$ denote a channel vector between the BS and a user terminal. Consider a space-time block code (STBC) coded transmission for common information broadcasting. Assume that the common information of binary bits is mapped to an STBC $\mathbf{S} \in \mathbb{C}^{M\times T}$ with $M \geq T$. This codeword matrix is then transmitted from the M antennas of the BS within T time slots. Then, the received signal is

$$[y_1, y_2, \ldots, y_T] = \mathbf{h}^{\mathbb{T}}\mathbf{S} + [z_1, z_2, \ldots, z_T] \tag{1}$$

where the channel $\mathbf{h}$ is assumed to keep constant within these T time slots, $^{\mathbb{T}}$ denotes the transpose, and $z_t \sim \mathcal{CN}(0, \sigma_{\mathrm{n}}^2)$ for $t = 1, 2, \ldots, T$ denotes the complex additive white Gaussian noise (AWGN).

Let $\mathbf{s}_t$ denote the tth column vector of the STBC $\mathbf{S}$. It is shown in [5]–[7] that the average received power[1] at time t is constant at M finite discrete angles if and only if all the elements in vector $\mathbf{F}_M\mathbf{s}_t$ have the same power, where $\mathbf{F}_M$ is the M-point DFT matrix. This motivates the following criterion [5]–[7].

Criterion of Omnidirectional Transmission at M Discrete Angles:

$$\left|\sum_{m=1}^{M} s_t(m)e^{-2\pi j(m-1)k/M}\right|^2 = c, \quad k = 0, 1, \ldots, M-1, \tag{2}$$

for some positive constant c, where $s_t(m)$ is the mth component of $\mathbf{s}_t$.

[1] In [4], the constant mean power across time t is considered, which makes the design problem much simpler.

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To have the transmission power efficiency and collect the spatial diversity order T, the following two criteria are imposed to the design of $\mathbf{S}$ [4]–[7].

Criterion of Constant Instant Transmission Power at Each Antenna: $|s_t(m)|^2 = c$, $m = 1, 2, \ldots, M$, for some positive constant c.

Criterion of Diversity Order T: The STBC $\mathbf{S}$ satisfies the full column rank, i.e., for any two distinct STBC codewords $\mathbf{S}_1$ and $\mathbf{S}_2$ of $\mathbf{S}$, the difference $\mathbf{S}_1 - \mathbf{S}_2$ has full column rank T.

Various systematic designs of STBC $\mathbf{S}$ to satisfy the above three criteria have been proposed in [5]–[7]. It should be emphasized that all the obtained designs only achieve the constant receiving power at M discrete angles. They may not have constant power at all analog angles ω, i.e.,

$$\left|\sum_{m=1}^{M} s_t(m)e^{-j(m-1)\omega}\right|^2 \neq c, \tag{3}$$

for any constant c.

In fact, it is not hard to see that as long as $M > 1$, the above property (3) always holds for any non-zero $s_t(m)$. This implies that it is impossible to achieve constant received power at instant time t at any angle. Thus, in order to consider every angle ω (not only the M discrete angles), we relax the requirement of constant received power at any instant time t to constant sum of received powers at two (or more) consecutive time slots as follows.

Criterion of Constant Received Sum Power at Any Angle: For all $\omega \in [0, 2\pi)$,

$$\left|\sum_{m=1}^{M} s_{2t-1}(m)e^{-j(m-1)\omega}\right|^2 + \left|\sum_{m=1}^{M} s_{2t}(m)e^{-j(m-1)\omega}\right|^2 = c, \tag{4}$$

for some positive constant c and $t = 1, 2, \ldots$, where, for simplicity, we only consider two consecutive time slots and a more general setting will be stated later. If we adopt the Fourier transform (FT) notation and let $S_t(\omega)$ denote the FT of sequence $s_t(m)$ (or vector $\mathbf{s}_t$), then, (4) can be equivalently re-written as

$$|S_{2t-1}(\omega)|^2 + |S_{2t}(\omega)|^2 = c, \quad \text{for all } \omega \in [0, 2\pi), \tag{5}$$

for some positive constant c.

One can see that as long as sequences $\mathbf{s}_{2t-1}$ and $\mathbf{s}_{2t}$ form a complementary sequence pair [8], they do satisfy the above criterion (4), while they may not satisfy the other two criteria of constant transmission power and full rank for a modulated STBC $\mathbf{S}$. The goal of this letter is to design an STBC $\mathbf{S}$ to satisfy full rank, constant transmission power, i.e., all elements in $\mathbf{S}$ have constant power, and constant received sum power (4).

III. Low Dimensional STBC and a Design

In a massive MIMO system, the number M of transmit antennas is large and it may cause a significant pilot overhead in downlink applications. A simple and well-known way to reduce the pilot overhead is to use a lower dimensional precoder $\mathbf{W}$ of size $M \times N$ in the transmission, where $1 < N << M$. In the STBC set up in this letter, it is the same as to use a lower dimensional STBC $\mathbf{S} = \mathbf{W}\mathbf{X}$, where the precoder $\mathbf{W}$ is pre-designed, independent of the channel or the information data, and $\mathbf{X}$ is an $N \times T$ STBC that is modulated by the information data. Since a fixed precoder $\mathbf{W}$ is independent of the channel, it can be absorbed to the channel at the receiver and the equivalent new channel becomes $\tilde{\mathbf{h}} = \mathbf{W}^{\mathbb{T}}\mathbf{h}$ that has size N by 1. Therefore, only N pilots are needed to estimate the equivalent channel at the receiver to detect the smaller size STBC $\mathbf{X}$. When N is much smaller than M, the pilot overhead can be much reduced. We next propose a particular design of $\mathbf{S} = \mathbf{W}\mathbf{X}$ to satisfy all the three criteria in Section II.

Let $\mathbf{a} = (a_1, \ldots, a_{M_0})^{\mathbb{T}}$ and $\mathbf{b} = (b_1, \ldots, b_{M_0})^{\mathbb{T}}$ be a pair of complementary sequences with constant module, i.e., $|a_m| = |b_m| = 1$ for all $m = 1, 2, \ldots, M_0$. Such complementary sequences can be polyphase complementray sequences and binary complementary sequences [8] of length M_0. Note that in this case, M_0 may not be arbitrarily chosen. We propose to use the following precoder $\mathbf{W}$ of size $2M_0 \times 2$, i.e., $M = 2M_0$ and $N = 2$:

$$\mathbf{W} = [\mathbf{w}_1, \mathbf{w}_2] = \begin{bmatrix} a_1 & 0 \\ 0 & b_1 \\ a_2 & 0 \\ 0 & b_2 \\ \vdots & \vdots \\ a_{M_0} & 0 \\ 0 & b_{M_0} \end{bmatrix}, \tag{6}$$

i.e., $w_1(2m-1) = a_m$ and $w_1(2m) = 0$, and $w_2(2m-1) = 0$ and $w_2(2m) = b_m$, for $m = 1, 2, \ldots, M_0$, where $\mathbf{w}_i = (w_i(1), \ldots, w_i(M))^{\mathbb{T}}$ for $i = 1, 2$. Note that since a length M_0 of a pair of polyphase complementary sequences may not be arbitrary, the number $M = 2M_0$ may not be arbitrary. A possible M_0 is a power of 2, i.e., $M_0 = 2^p$ for a positive integer p, and then, $M = 2^{p+1}$ for a positive integer p. For more detailed constructions, we refer the reader to [8] as well as the recent literatures on the constructions of polyphase complementary sequences. From (6), one can see that the two columns of the precoder $\mathbf{W}$ are orthogonal each other, which is a desired property for a precoder in signal processing, such as it may maintain the transmission signal power etc.

We propose to use the 2×2 Alamouti code as the smaller size STBC $\mathbf{X}$, [3]:

$$\mathbf{X} = \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix} \tag{7}$$

where x_1 and x_2 take PSK signals independently with $|x_1| = |x_2| = 1/\sqrt{2}$.

We now want to show that the STBC $\mathbf{S} = \mathbf{W}\mathbf{X}$ with the above $\mathbf{W}$ in (6) and $\mathbf{X}$ in (7) satisfies all the three criteria proposed in the preceding section. Let us first prove that it satisfies the constant received sum power criterion. Since $\mathbf{a}$ and $\mathbf{b}$ are a pair of complementary sequences, their FT satisfy

$$|A(\omega)|^2 + |B(\omega)|^2 = c, \quad \text{for all } \omega \in [0, 2\pi), \tag{8}$$

for some positive constant c. From the construction of $\mathbf{W}$ in (6), it is not hard to see that $W_1(\omega) = A(2\omega)$ and

$W_2(\omega) = e^{-j\omega}B(2\omega)$ where $W_i(\omega)$ is the FT of $\mathbf{w}_i$ for $i = 1, 2$. For the STBC $\mathbf{S} = [\mathbf{s}_1, \mathbf{s}_2] = \mathbf{WX}$, from (6) and (7) we have

$$\mathbf{S} = [\mathbf{s}_1, \mathbf{s}_2] = \begin{bmatrix} a_1x_1 & a_1x_2 \\ -b_1x_2^* & b_1x_1^* \\ a_2x_1 & a_2x_2 \\ -b_2x_2^* & b_2x_1^* \\ \vdots & \vdots \\ a_{M_0}x_1 & a_{M_0}x_2 \\ -b_{M_0}x_2^* & b_{M_0}x_1^* \end{bmatrix}. \tag{9}$$

Thus, it is not hard to see that

$$[S_1(\omega), S_2(\omega)] = [A(2\omega), e^{-j\omega}B(2\omega)]\mathbf{X}, \tag{10}$$

where $\mathbf{X}$ is the Alamouti code in (7). Thus,

$$\begin{aligned} &[S_1(\omega), S_2(\omega)][S_1(\omega), S_2(\omega)]^{\mathbb{H}} \\ &= [A(2\omega), e^{-j\omega}B(2\omega)]\mathbf{X}([A(2\omega), e^{-j\omega}B(2\omega)]\mathbf{X})^{\mathbb{H}} \\ &= [A(2\omega), e^{-j\omega}B(2\omega)]\mathbf{X}\mathbf{X}^{\mathbb{H}}[A(2\omega), e^{-j\omega}B(2\omega)]^{\mathbb{H}} \\ &= [A(2\omega), e^{-j\omega}B(2\omega)][A(2\omega), e^{-j\omega}B(2\omega)]^{\mathbb{H}} \quad (11) \\ &= |A(2\omega)|^2 + |B(2\omega)|^2, \quad (12) \end{aligned}$$

where (11) is due to the Alamouti code structure of $\mathbf{X}$ in (7) with PSK signals x_1 and x_2, and $\mathbb{H}$ stands for the conjugate transpose.

Due to the complementary sequence property (8) of $\mathbf{a}$ and $\mathbf{b}$, the above identity (12) implies $|S_1(\omega)|^2+|S_2(\omega)|^2 = c$ for a positive constant c for all angle $\omega \in [0, 2\pi)$, i.e., the constant received sum power property (5) (or (4)) is proved.

From (9) and $|x_1| = |x_2|$ and $|a_1| = \cdots = |a_{M_0}| = |b_1| = \cdots = |b_{M_0}|$, the constant transmission power criterion of $\mathbf{s}_i$ for $i = 1, 2$ is satisfied.

Since the Alamouti code $\mathbf{X}$ in (7) has the full rank property, it is clear that the low dimensional STBC $\mathbf{S} = \mathbf{WX}$ has diversity order $T = 2$, i.e., the full column rank or the full low dimensional spatial diversity, is proved as well.

In summary, the above low dimensional (or precoded) STBC design $\mathbf{S}$ does satisfy all the criteria presented in Section II.

Furthermore, since the two information symbols x_1 and x_2 in the Alamouti code $\mathbf{X}$ are independent, the fast symbol-wise maximum-likelihood (ML) decoding of the Alamouti code still holds. This is different from the designs in [5]–[7] to have constant received power at any instant time t for finitely many discrete angles, where the two information symbols x_1 and x_2 may need to be dependent in general.

As a side result, the two sequences $\mathbf{s}_1$ and $\mathbf{s}_2$ in (9) form another pair of complementary sequences from the above proof. This provides a way to construct new pairs of complementary sequences with any constant module symbols x_1 and x_2 (or any 2×2 unitary matrix $\mathbf{X}$ in $[\mathbf{s}_1, \mathbf{s}_2] = \mathbf{S} = \mathbf{WX}$ with $\mathbf{W}$ in (6)) from a known pair of complementary sequences $\mathbf{a}$ and $\mathbf{b}$, where the sequence length is doubled. It is obvious that $[1, 1]^{\mathbb{T}}$ and $[1, -1]^{\mathbb{T}}$ form a pair of complementary sequences of length 2. From this pair, one will be able to construct longer ones of length 2^p for any $p > 1$ following the above method.

IV. Discussions and Generations on Constant Received Sum Power

Another systematic method to construct complex-valued complementary sequences is to use polyphase[2] sequences/components of a sequence. Then, complementary sequences $\mathbf{s}_1$ and $\mathbf{s}_2$ in $\mathbf{S}$ (or having the constant received sum power property) can be constructed through lossless matrices (and/or paraunitary matrices) [9] ([10] for a shorter reference). Vector (or sequence) $\mathbf{s}_t$ for each t, $t = 1, 2$, is split into two subvectors (or subsequences) $s_{t,0}(m)$ and $s_{t,1}(m)$ of even and odd indices, respectively, called two polyphase components of $\mathbf{s}_t$ [9]. Note that general T polyphase components $s_{t,n}(m)$, $0 \le n \le T-1$, of $\mathbf{s}_t$ can be similarly obtained by evenly splitting it into T subvectors (or subsequences). In terms of their FT, $S_{t,n}(\omega)$ is the nth polyphase component of $S_t(\omega)$. Then, the constant received sum power (4) or (5) is ensured by the following lossless matrix property [9]:

$$\boldsymbol{\mathcal{S}}^{\mathbb{H}}(\omega)\boldsymbol{\mathcal{S}}(\omega) = c\mathbf{I}_2, \text{ for all } \omega \in [0, 2\pi), \tag{13}$$

for some positive constant c, where

$$\boldsymbol{\mathcal{S}}(\omega) = \begin{bmatrix} S_{2t-1,0}(\omega) & S_{2t,0}(\omega) \\ S_{2t-1,1}(\omega) & S_{2t,1}(\omega) \end{bmatrix}, \tag{14}$$

$\mathbf{I}_2$ is the 2 by 2 identity matrix. $\boldsymbol{\mathcal{S}}(\omega)$ is called a polyphase matrix of $S_{2t-1}(\omega)$ and $S_{2t}(\omega)$ (or $\mathbf{s}_{2t-1}$ and $\mathbf{s}_{2t}$). Therefore, one can see that the construction of vectors $\mathbf{s}_{2t-1}$ and $\mathbf{s}_{2t}$ that satisfy the constant received sum power criterion (4) or (5) can be obtained by constructing a lossless matrix $\boldsymbol{\mathcal{S}}(\omega)$ in (13) as a polyphase matrix, which has complete factorization and systematic constructions [9] and [11] shown below in general forms. Unfortunately, the design presented in Section III does not satisfy the above lossless matrix property (13). $S_1(\omega)$ and $S_2(\omega)$ from the STBC $\mathbf{S} = \mathbf{WX}$ in Section III have the following polyphase matrix:

$$\boldsymbol{\mathcal{S}}(\omega) = \mathbf{X}\begin{bmatrix} A(\omega) & 0 \\ 0 & B(\omega) \end{bmatrix}.$$

Then,

$$\boldsymbol{\mathcal{S}}^{\mathbb{H}}(\omega)\boldsymbol{\mathcal{S}}(\omega) = \begin{bmatrix} |A(\omega)|^2 & 0 \\ 0 & |B(\omega)|^2 \end{bmatrix} \ne c\mathbf{I}_2$$

for any constant c. This means that the above lossless polyphase matrix construction is only sufficient but not necessary. The reason is as follows. The lossless matrix property (13) is equivalent to the lossless matrix property of the following matrix (setting $t = 1$ in (5) and (13)) [9]:

$$\begin{bmatrix} S_1(\omega) & S_2(\omega) \\ S_1(-\omega) & S_2(-\omega) \end{bmatrix},$$

which means that $[S_1(\omega), S_2(\omega)]$ and $[S_1(-\omega), S_2(-\omega)]$ are orthogonal to each other, in addition to the complementary sequence property of $S_1(\omega)$ and $S_2(\omega)$.

[2]The word "polyphase" here has different meaning from that in "polyphase complementary sequences". The later means multiple phases.

Before we describe the general construction, let us first generalize the constant received sum power criterion (4). As we have explained earlier, to have a constant received power from any angle at an instant time t is impossible. The minimum number of time slots to satisfy constant received sum power is 2 as (4). It can be generalized to T time slots for any positive integer T with $T \geq 2$:

$$\sum_{t=1}^{T}\left|\sum_{m=1}^{M} s_t(m)e^{-j(m-1)\omega}\right|^2 = c \text{ for all } \omega \in [0, 2\pi), \tag{15}$$

for some positive constant c. Its FT domain representation is

$$\sum_{t=1}^{T}|S_t(\omega)|^2 = c \text{ for all } \omega \in [0, 2\pi), \tag{16}$$

for some positive constant c. Similar to the above $T = 2$ case, The constant received sum power criterion (15) or (16) can be guaranteed by constructing a lossless polyphase matrix $\boldsymbol{\mathcal{S}}(\omega)$ of T by T, [9],

$$\boldsymbol{\mathcal{S}}^{\mathbb{H}}(\omega)\boldsymbol{\mathcal{S}}(\omega) = c\mathbf{I}_T, \text{ for all } \omega \in [0, 2\pi), \tag{17}$$

for some positive constant c, where $\mathbf{I}_T$ is the T by T identity matrix and

$$\boldsymbol{\mathcal{S}}(\omega) = \begin{bmatrix} S_{1,0}(\omega) & \cdots & S_{T,0}(\omega) \\ \vdots & \vdots & \vdots \\ S_{1,T-1}(\omega) & \cdots & S_{T,T-1}(\omega) \end{bmatrix} \tag{18}$$

where $S_{t,n}(\omega)$, $0 \leq n \leq T-1$, are the polyphase components of $S_t(\omega)$ (or $\mathbf{s}_t$) for $1 \leq t \leq T$.

A lossless matrix $\boldsymbol{\mathcal{S}}(\omega)$ in (13) and (14) or (18) can be completely (necessarily and sufficiently) factorized and systematically constructed via, see [9],

$$\boldsymbol{\mathcal{S}}(\omega) = \prod_{l=1}^{L}\mathbf{V}_l(\omega)\mathbf{V} \tag{19}$$

for some integer $L \geq 0$, where when $L = 0$, it means that the term $\mathbf{V}_l(\omega)$ does not appear in (19), $\mathbf{V}$ is a $T \times T$ constant unitary matrix, and

$$\mathbf{V}_l(\omega) = \mathbf{I}_T - \mathbf{v}_l\mathbf{v}_l^{\mathbb{H}} + e^{-j\omega}\mathbf{v}_l\mathbf{v}_l^{\mathbb{H}}, \tag{20}$$

where $\mathbf{v}_l \in \mathbb{C}^{T\times 1}$ is a T by 1 constant column vector of unit norm.

From the above factorization, one can construct a lossless matrix $\boldsymbol{\mathcal{S}}(\omega)$ by arbitrarily taking L many T by 1 vectors $\mathbf{v}_l$ of unit norm and also a T by T unitary matrix $\mathbf{V}$ and putting them into (19)-(20). The good property about the factorization (19) is that the unit norm vectors $\mathbf{v}_l$ in the degree one construction are almost free and have only one constraint, i.e., the unit norm constraint. However, it does not guarantee the length of each polyphase component $S_{t,n}(\omega)$, i.e., it does not guarantee the length of each $S_t(\omega)$. Since the length of $S_t(\omega)$ is M, the number of transmit antennas, and furthermore, its coefficients $s_t(m)$ need to have constant module, it is critically important to control the length (or the order) of each $S_{t,n}(\omega)$. A complete factorization in terms of the order (or the length) of $S_{t,n}(\omega)$ is obtained in [11] as follows. Let $K = M/T$ be an integer. Then, the length of each $S_{t,n}(\omega)$ is K (or its order is $K-1$) and a lossless matrix $\boldsymbol{\mathcal{S}}(\omega)$ of order $K-1$ has the following order one factorization, [11]:

$$\boldsymbol{\mathcal{S}}(\omega) = \mathbf{U}_{K-1}\mathbf{\Lambda}_{K-1}(\omega)\mathbf{U}_{K-2}\mathbf{\Lambda}_{K-2}(\omega)\cdots\mathbf{U}_1\mathbf{\Lambda}_1(\omega)\mathbf{U}_0, \tag{21}$$

where $\mathbf{U}_k$ are some $T \times T$ unitary matrices, and

$$\mathbf{\Lambda}_k(\omega) = \begin{bmatrix} \mathbf{I}_{T-r_k} & \mathbf{0} \\ \mathbf{0} & e^{-j\omega}\mathbf{I}_{r_k} \end{bmatrix}, \tag{22}$$

where r_k are some positive integers with $1 \leq r_k \leq T-1$ for $0 \leq k \leq K-1$. The above factorization is also necessary and sufficient [11] and can be used to construct a lossless matrix of order $K = M/T$. In addition to the constant module property of all the coefficients of $\mathbf{s}_t$ for every t, the STBC $\mathbf{S} = [\mathbf{s}_1, \cdots, \mathbf{s}_T]$ formed by all these sequences $\mathbf{s}_t$ needs to satisfy the full rank property to achieve the spatial diversity of order T as the third criterion in the preceeding section. How to design them is still open.

V. Conclusion

In this letter, we have proposed a space-time block code design for omindirectional transmission in a downlink massive MIMO system. It achieves 1) the constant sum of the received signal powers in two consecutive time slots at any angle (not just at finite discrete angles as what was previously down); 2) constant signal power at any transmit antenna at any instant time; 3) diversity order 2, i.e., full column rank of the low dimensional STBC. Furthermore, it has the fast symbol-wise ML decoding. We have re-formulated the design problem differently using lossless polyphase matrices but its concrete design is still open.

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