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Mutual Coupling Reduction for Dual-Band MIMO Antenna via Artificial Transmission Line

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SUMMARY A dual-band decoupling strategy via artificial transmission line (TL) for closely spaced two-element multiple-input multiple-output antenna (MIMO) is proposed, which consists of two composite right-/left-handed TLs for dual-band phase shifting and a cross-shaped TL for susceptibility elimination to counteract the real and imaginary part of the mutual coupling coefficient $S_{21}$ at dual frequency bands, respectively. The decoupling principle and detailed design process of the dual-band decoupling scheme are presented. To validate the dual-band decoupling technique, a closely spaced dual-band MIMO antenna for 5G (sub-6G frequency band) utilization is designed, fabricated, and tested. The experimental results agree well with the simulation ones. A dual-band of 3.40 GHz–3.59 GHz and 4.79 GHz–4.99 GHz ($S_{11} < -10$ dB, $S_{21} < -20$ dB) has been achieved, and the mutual coupling coefficient $S_{21}$ is significantly reduced to 12 dB and 16.1 dB at 3.5 GHz and 4.9 GHz, respectively. In addition, the proposed dual-band decoupling scheme is antenna independent, and it is very suitable for other tightly coupled dual-band MIMO antennas.

key words: Composite right-/left-handed transmission line, Cross-shaped TL, Dual-band decoupling, Multiple-input multiple-output (MIMO), Mutual coupling.

1. Introduction

Multiple-input multiple-output (MIMO) antenna technique exhibits fairly good properties with high data throughput and low time delay, which has been universally adopted in 5G and radar systems[1]. However, the miniaturization trend in modern wireless systems making the antenna design still needs to consider the limited space available for mounting multiple antennas. Severe electromagnetic (EM) coupling would occur if the inter-element distance is less than half of the operating wavelength in free space. The strong EM coupling not only has effects on the radiation efficiency of the MIMO antenna but also declines the working capability in engineering applications for the MIMO antenna.

To solve the technical contradiction mentioned above, several methodologies have been presented for EM coupling reduction. Neutralization-line technique[2], [3] is a common decoupling approach, which aims to generate an opposite coupling current to cancel the mutual coupling for the MIMO antenna. Although the neutralization-line technique can reduce EM coupling, to the best of the author’s knowledge this way is essentially based on experimental attempts and it is difficult to find a specific design criteria to use now.

Defected ground structure (DGS)[4], [5] can also be used to increase the isolation between MIMO antenna elements. However, the DGS is typically etched from the ground plane, it will destroy the integrity of the ground plane and not benefit the back-radiation performance of the decoupled MIMO antennas.

Metamaterial (MTM) is an engineered structure and possesses many interesting EM characteristics, such as negative permittivity or negative permeability[6], [7]. Electromagnetic bandgap structures (EBG)[8], [9], MTM-photonic bandgap[10], capacitively-loaded loop[11],[12], and metasurface[13], [14] are different formats of MTM. In virtue of the extraordinary characteristics of MTM, the propagation of the surface wave can be prevented to alleviate the EM coupling among the adjacent antenna elements[8]–[15]. However, the design process of MTM is relatively complex and it always occupies a large volume which will dramatically increase the package size of the communication system.

The decoupling and matching network (DMN)[16]–[19] can furnish an effective and systematic way to mitigate the EM coupling and improve the isolation in an array of MIMO systems. The working mechanism of the DMN is to remove the real part of the coupling coefficient $S_{21}$ by utilizing a section of transmission line (TL), and then, a well designed anti-susceptance is designed to vanish the imaginary part of the coupling coefficient and enhance the isolation for the coupled MIMO antenna[19].

However, the traditional DMNs[18]–[20] are limited in single-band decoupling due to the linear phase response of the TL, which is not consistent with the trend of dual-band or multi-band wireless systems. In order to tackle this problem, some dual-band decoupling methods have been reported in [21]–[25], such as the dual-band rat-race coupler in[22], the multipath decoupling circuit in[24], and the hybrid resonant structure in[25], et al. Though the aforementioned dual-band decoupling scheme achieves a good mutual coupling reduction, some problems still exist, for instance, the narrow operating band or the center-to-center space between antenna elements is actually large.

In this work, a dual-band decoupling strategy is presented and investigated for the closely spaced two-element dual-band MIMO antenna, which consists of two composite
right/left-handed (CRLH) TLs and a cross-shaped TL. The CRLH-TL with nonlinear phase response is utilized for dual-band phase shifting to eliminate the real part of the coupling coefficient $S_{21}$. After that, a cross-shaped TL is employed to counteract the imaginary part of $S_{21}$ at dual frequency band, then to decouple the dual-element MIMO antenna. The decoupling principle and design process of this methodology are given in Section 2. To validate the proposed decoupling scheme, a closely spaced two-element MIMO antenna for 5G (sub-6G) utilization is designed, fabricated, and analyzed in Section 3. In final, the conclusions are drawn in Section 4.

2. Dual-band Decoupling Scheme and Theory Analysis

2.1 The Proposed Dual-band Decoupling Scheme

As is known, for a closely spaced two-element dual-band MIMO antenna given in Fig. 1, impedance matching is much easier than decoupling. Assuming the the closely spaced two-element MIMO antenna matches their port input impedance $Z_0$ very well at the reference plane $T1$ but suffers severe EM coupling. Then, the coupled S-parameter matrix of the two-port network for the two-element MIMO antenna at the reference plane $T1$ can be approximately expressed as

$$\begin{bmatrix} S_{11}^{T1} & S_{12}^{T1} \\ S_{21}^{T1} & S_{22}^{T1} \end{bmatrix} = \begin{bmatrix} 0 & e^{j\theta_L}\phi_T^{T1}(\omega_L) \\ e^{-j\theta_L}\phi_T^{T1}(\omega_L) & 0 \end{bmatrix}, \quad (1)$$

where $|S_{12}^{T1}(\omega)|$ and $|S_{21}^{T1}(\omega)|$ denote the magnitude, while $\phi_T^{T1}(\omega)$ represents the phase of insertion loss of the two port network at $T1$ plane. Typically, the insertion loss $|S_{12}^{T1}| = S_{21}$ stands for the levels of the EM mutual coupling between two MIMO antenna elements, with a higher value of $S_{21}$ indicating a stronger degree of coupling.

As depicted in Fig. 1, the dual-band decoupling scheme is composed of two CRLH-TLs and a cross-shaped TL to eliminate the real and imaginary part of the mutual coupling coefficient $S_{21}$, respectively, at the lower resonance frequency $\omega_L$ and upper resonance frequency $\omega_H$ ($\omega_L < \omega_H$).

To eliminate the real part of the $S_{21}$, the characteristic impedance of the CRLH-TL added at the $T1$ plane should keep the same with the port input impedance $Z_0$ and according to the TL theory [26], the electric length of the CRLH-TL at the two resonance frequencies should follow as:

$$\theta_L = \frac{1}{2} \left[ \phi_T^{T1}(\omega_L) \pm \pi \right] , \quad k \in Z, \quad (2a)$$

where $Z$ denotes an integer set, $\theta_L$ and $\theta_H$ represent the electric length of the CRLH-TL at two resonance frequencies $\omega_L$ and $\omega_H$ respectively. After that, the S-parameter matrix of the two-port network at reference plane $T2$ can be converted into the corresponding admittance matrix \(Y_{T2}\) [26], that is

$$\begin{bmatrix} y_{11}^{T2} & y_{12}^{T2} \\ y_{21}^{T2} & y_{22}^{T2} \end{bmatrix} = \frac{1}{Z_0} \begin{bmatrix} 1 + |S_{12}^{T1}(\omega_L)|^2 & j2\pi |S_{21}^{T1}(\omega_L)| \\ j2\pi |S_{12}^{T1}(\omega_L)| & 1 + |S_{21}^{T1}(\omega_L)|^2 \end{bmatrix} \cdot \begin{bmatrix} 2 \chi_1^2 & 2 \chi_1 \chi_2 \\ 2 \chi_1 \chi_2 & 2 \chi_2^2 \end{bmatrix}, \quad (3a)$$

where $\chi_1 = \frac{1}{2} \left[ \phi_T^{T1}(\omega_H) \pm \pi \right]$.

$$\begin{bmatrix} y_{11}^{T2} & y_{12}^{T2} \\ y_{21}^{T2} & y_{22}^{T2} \end{bmatrix} = \frac{1}{Z_0} \begin{bmatrix} 1 + |S_{12}^{T1}(\omega_H)|^2 & j2\pi |S_{21}^{T1}(\omega_H)| \\ j2\pi |S_{12}^{T1}(\omega_H)| & 1 + |S_{21}^{T1}(\omega_H)|^2 \end{bmatrix} \cdot \begin{bmatrix} 2 \chi_2^2 & 2 \chi_1 \chi_2 \\ 2 \chi_1 \chi_2 & 2 \chi_1^2 \end{bmatrix} \cdot \begin{bmatrix} 2 \chi_2^2 & 2 \chi_1 \chi_2 \\ 2 \chi_1 \chi_2 & 2 \chi_1^2 \end{bmatrix}, \quad (3b)$$

Once the real part of the $S_{21}$ is vanished, as illustrated in Fig. 1, the cross-shaped TL is in shunted at the reference plane $T2$. It will provide two reactive elements of susceptance $b_L$ and $b_H$ to eliminate the imaginary part of $S_{21}$ at two resonance frequencies $\omega_L$ and $\omega_H$, respectively. The susceptances $b_L$ and $b_H$ should be equal to the inverse of the transfer admittance of $Y_{T2}$ at two resonance frequencies $\omega_L$ and $\omega_H$, namely

$$b_L = \frac{1}{Z_0} \frac{\pm 2 |S_{12}^{T1}(\omega_L)|}{\pm 2 |S_{21}^{T1}(\omega_L)|} , \quad b_H = \frac{1}{Z_0} \frac{\pm 2 |S_{12}^{T1}(\omega_H)|}{\pm 2 |S_{21}^{T1}(\omega_H)|}, \quad (4)$$

where $\pm 2 |S_{12}^{T1}(\omega)|$ and $\pm 2 |S_{21}^{T1}(\omega)|$ represent the electric length of the CRLH-TL for dual-band phase shifting to eliminate the real part of $S_{21}$ at the lower and upper bands. According to [27], the unit phase responses for the RH-TL and LH-TL are

$$\phi_{L_{\text{unit}}} = - \arctan \left[ \frac{C_L Z_0 L_L - \frac{1}{2} \omega^2 C_L Z_0 L_L}{1 - 2\omega^2 C_L L_L} \right], \quad (5a)$$

$$\phi_{R_{\text{unit}}} = - \arctan \left[ \frac{C_R Z_0 L_L - \frac{1}{2} \omega^2 C_R Z_0 L_L}{1 - 2\omega^2 C_R L_L} \right], \quad (5b)$$

The CRLH-TL is employed for dual-band phase shifting to eliminate the real part of $S_{21}$ and whose equivalent circuit model is shown in Fig. 2(a) [22], [27], [28]. With respect to the LH part of the CRLH-TL, which is formed by two series capacitors ($2C_L$) along with a shunt inductor ($L_L$) inserted in-between. Because the phase response of LH-TL is positive (phase advance), it is utilized to combine the RH-TL (phase lag) to exhibit nonlinear phase response at two desired frequency bands. Then, the CRLH-TL is employed for dual-band phase shifting to eliminate the real part of $S_{21}$ at the lower and upper bands. According to [27], the unit phase responses for the RH-TL and LH-TL are
where $Z_{0L}$ and $Z_{0R}$ represent the characteristic impedance of the RH-TL and LH-TL, respectively, given by

$$Z_{0L} = \sqrt{L_L/C_L}, Z_{0R} = \sqrt{L_R/C_R}. \tag{6}$$

Owing to the CRLH-TL in series combined with the RH-TL and LH-TL, the phase of CRLH-TL can be worked out as

$$\phi_C = N\varphi_{\text{unit}}^L + N\varphi_{\text{unit}}^R, N \in \mathbb{Z}. \tag{7}$$

Solving Eq. (5) is relatively complex. But when the phase response of the CRLH-TL unit is much smaller than $\pi/2$, it can be simplified as

$$\varphi_{\text{unit}}^L = \left( \omega \sqrt{L_L C_L} \right)^{-1}, \varphi_{\text{unit}}^R = -\omega \sqrt{L_R C_R}. \tag{8}$$

In order to eliminate the real part of $S_{21}$ while keeping the impedance matching at two resonance frequencies $\omega_L$ and $\omega_H$ simultaneously. The characteristic impedance of the CRLH-TL should keep the same with the port input impedance $Z_0$ at the reference plane T1 ($Z_{0R} = Z_{0L} = Z_0$), and the phase response of the CRLH-TL should surrender to Eq. (2), that is

$$\phi_C(\omega_L) = N \left[ \frac{1}{\omega_L \sqrt{L_L C_L}} - \omega_L \sqrt{L_R C_R} \right] = -\theta_L. \tag{9a}$$

$$\phi_C(\omega_H) = N \left[ \frac{1}{\omega_H \sqrt{L_L C_L}} - \omega_H \sqrt{L_R C_R} \right] = -\theta_H. \tag{9b}$$

When the S-parameter matrix $[S]$ of a MIMO antenna array is obtained, for a given number $N$, combined with Eqs. (6)–(9), the values for $L_L$, $C_L$, $L_R$, and $C_R$ of the CRLH-TL unit can be fetched. As depicted in Fig. 2(a), two conventional microstrip lines on each side of two series LH-TL units are employed to form the RH part of the CRLH-TL, whose unit electrical length $\theta_R$ (at $\omega_L$) can be calculated using Eq. 8 regarding about the unit phase response of RH-TL ($\theta_R = -N\varphi_{\text{unit}}^R$).

2.2 The CRLH-TL for Dual-band Phase Shifting

2.3 Cross-shaped TL for Susceptance Elimination

After the designed CRLH-TL is connected to the input port at the reference plane T1, the coupling coefficient $S_{21}$ becomes pure imaginary at the dual frequency bands, which can be eliminated by dual-band susceptance $jB$ as shown in Fig. 1.

The cross-shaped TL can provide dual-band susceptance[29], whose geometry and parameters are illustrated in Fig. 2(b). It is employed to wipe off the imaginary part of $S_{21}$. To be more precise, the transfer admittance $Y_{21}$ of the cross-shaped TL is set to equal to $jB_L$ and $jB_H$ at two resonance frequencies $\omega_L$ and $\omega_H$ respectively, to decouple the dual-band MIMO antenna. Though it will produce two transcendental equations that are difficult to solve, the parameters of the cross-shaped TL can be optimized by Keysight Advanced Design System (ADS) to acquire an iteration solution. Considering that the self admittance $Y_{11}$ is not strictly equal to $1/Z_0$ after the cross-shaped TL in shunted at T2 plane, a simple dual-band impedance matching circuit is used at reference plane T3, as shown in Fig. 1.

3. Experimental Verification

In the following section, a closely spaced two-element dual-band MIMO antenna operating at 3.5 GHz and 4.9 GHz is designed to validate the proposed dual-band decoupling mechanism. All circuits are printed on a low-cost 0.8-mm-thick FR4 substrate with relative dielectric constant $\varepsilon_r = 4.4$ and loss tangent $\tan \delta = 0.02$.

3.1 Closely Spaced Monopole MIMO Antenna

Fig. 3 depicts the layout and detailed dimensions for the closely spaced two-element dual-band monopole MIMO antenna (renamed reference MIMO antenna here for easy description in the following article). A meander line inductor paralleled with two interdigital capacitors forms an LC parallel resonant circuit, which is inserted at $l_{a1}$ in each monopole.
antenna element to realize dual-band. When the LC parallel circuit resonances at \( \omega_1 \) (4.9 GHz), the equivalent circuit is open, leading to the stripline \( l_{1l} \) dominating the radiation at the higher resonance frequency. With respect to the radiation in the lower frequency band, the whole monopole stripline \( l_{2l} \) will take responsibility for radiation. Therefore, by properly adjusting the lengths of \( l_{1l} \) and \( l_{2l} \) to one-quarter wavelength at \( \omega_1 \) and \( \omega_L \) (3.5 GHz), respectively, while concurrently calibrating the resonant frequency of the LC parallel circuit to \( \omega_1 \), a desired dual-band monopole MIMO antenna can be realized.

In this work, all parameters are simulated and optimized in 3-D full wave simulation software ANSYS HFSS. The simulated S-parameters of the reference MIMO antenna are shown in Fig. 4(a). At the lower frequency band, the -10 dB impedance bandwidth (BW) is 480 MHz (from 3.27 GHz to 3.75 GHz), while at the upper band, it is 560 MHz (from 4.66 GHz to 5.22 GHz). In addition, Fig. 4(b) and (c) also illustrate the current distribution on the monopole and its related 3D radiation pattern. It is found most of the current at 3.5 GHz propagates along stripline \( l_{1l} \), meander line inductor, and then reaches the end of the monopole, which will precipitate the radiation at the lower frequency band. While at 4.9 GHz, two null current points appear at Null 1 and Null 2, which confirms the resonance in the LC parallel circuit, consequently, stripline \( l_{2l} \) will dominate the radiation at the upper frequency band.

Since the center-to-center distance (CCD) of two antenna elements is only 9.5 mm, which is about 0.11 \( \lambda_L \) \( (\lambda_L \) represents the free-space wavelength at 3.5 GHz), it will make the antenna elements suffer severe EM coupling. As depicted in Fig. 4, the simulated mutual coupling coefficient \( S_{21} \) reaches up to -8.54 dB and -10.39 dB at 3.5 GHz and 4.9 GHz, respectively. Evidently, the isolation for the reference MIMO antenna is not acceptable, therefore, much attention should be paid to the EM coupling reduction among the adjacent antenna elements, and a dual-band decoupling strategy is desperately required.

### 3.2 Phase Shifting and Susceptance Elimination for \( S_{21} \)

For a given closely spaced two-element monopole MIMO antenna, the S-parameters matrix \( [Z^{T1}] \) at the reference plane T1 is defined. The CRLH-TL with characteristic impedance \( Z_0 \) is added at reference plane T1 to realize phase shifting and eliminate the real part of \( S_{21} \). As shown in Fig. 5(a), the phase of \( S_{21} \) for the reference MIMO antenna without adding CRLH-TL at T1 plane is -102.87° and -109.31° at 3.5 GHz and 4.9 GHz, respectively. According to Eq. (2), the electric length of the CRLH-TL (\( \theta_1 \) and \( \theta_H \)) for phase shifting can be obtained, and it is 83.56° and 170.35° when \( k \) is chosen to be 1 and 2 at Eq. 2(a) and Eq. 2(b), respectively. Following the design rule proposed in Subsection 2.2, the circuit parameters of the CRLH-TL \( (L_1, C_1, \text{and} \theta_H) \) depicted in Fig. 2(a) can be calculated from Eqs. (6)–(9), which are 3.35 nH, 1.34 pF, and 80.68°.

Fig. 5(a) reveals the phase response of the CRLH-TL simulated by the Keysight ADS. The simulation values at two resonance frequencies \( \omega_L \) (3.5 GHz) and \( \omega_H \) (4.9 GHz) are -83.5° and -170.3°, which closely match the calculated values by Eq. (2). Though the circuit values in ADS have been slightly tuned to 3.4 nH, 1.4 pF, and 80.18° for realizing consideration. The phase response of \( S_{21} \) of the reference MIMO antenna with adding two CRLH-TLs at T2 plane is also given in Fig. 5(a), and it is 90.46° and -89.89°, respectively when it operates at 3.5 GHz and 4.9 GHz, which means the real part of \( S_{21} \) is vanished after the dual-band phase shifting. Fig. 5(b) shows the effects on the S-parameters before and after phase shifting for \( S_{21} \), it is found that there is no important influence on the impedance bandwidth and mutual coupling coefficient \( S_{21} \). Though a slightly offsetting at the low band is found, the -10 dB impedance BW still covers from 3.33 GHz to 3.74 GHz.

Once the real part of \( S_{21} \) is dispelled, as depicted in Fig. 6(a), the transfer admittance \( Y_{21} \) of the reference MIMO

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**Fig. 5** (a) Phase of \( S_{21} \) for the reference MIMO antenna with and without adding CRLH-TL at T1 plane and the phase of \( S_{21} \) for CRLH-TL itself. (b) S-parameters of the reference MIMO antenna with and without CRLH-TL integration at T1 plane.

**Fig. 6** (a) Simulated \( Y_{21} \) of the cross-shaped TL and those of the reference MIMO antenna with CRLH-TL integration at T2 plane. Simulated (b) S-parameters, (c) Real and (d) Imaginary part of \( Z_{21} \) with different \( l_{inl} \) during the impedance matching process.
antenna at T2 plane are $0 - j0.0141$S and $0 + j0.0113$ S at
3.5 GHz and 4.9 GHz respectively. The cross-shaped TL
is considered to provide the inverse of the transfer admittance
$Y_{21}$ to wipe off the imaginary part of $Y_{21}$ and decouple
the reference MIMO antenna at dual-frequency band. As
mentioned in Subsection 2.3, it is difficult to obtain an an-
alytic solution for Eq. (4). But the function of optimization
in Keysight ADS can provide an iterative solution, and
the parameters of the cross-shaped TL iterated by ADS are
$0.8(w_1), 2.2(w_2), 1.3(w_3), 9.4(l_1), 18.1(l_2), 11.3(l_2), \text{(unit: mm)}$.
However, the $Y_{21}$ obtained by ADS at 3.5 GHz and
4.9 GHz are $+j0.07$ S ($jB_1$) and $-j0.011$ S ($jB_2$), which is not
good enough to provide the anti-susceptance at the lower fre-
quency band. To make the imaginary part of $Y_{21}$ as closely
as possible to $+j0.0141$S and $-j0.0113$ S at 3.5 GHz and
4.9 GHz. Further optimization of the cross-shaped TL is
conducted in ANSYS HFSS, as shown in Fig. 6(a), the final
optimized $Y_{21}$ values at those frequencies are $+j0.0133$ S and
$-j0.0096$ S. The layout and parameters of the cross-shaped TL
utilized in HFSS are depicted in Fig. 7(a).

After that, the dual-band reference MIMO antenna is
decoupled and the related S-parameters at T3 plane is illus-
trated in Fig. 6(b) (gray dash dot dot line line). It can be
seen the resonance point at the lower band has been slightly
shifted. In addition, the resonance dip observed at $S_{11}$ below
-10 dB around the 4.25 GHz band primarily originates the
$Y_{12}$ of cross-shaped TL, but the peak gain of the decoupled
MIMO antenna in this band is very low and the filter among
the transceiver system and power amplifier will guarantee
the undesired signal flow to communication system. To re-
alize the input impedance matches at dual-frequency band,
the grounded stub with a length of $l_{m2}$ is inserted in-between
a TL to adjust the input impedance of the decoupled MIMO
antenna at T3 plane. Fig. 6(c) and (d) illustrate the effects
of $l_{m2}$ (from 11 to 13.4 mm) on the real and imaginary of
the input impedance. When $l_{m2}$ was tuned to 11.8 mm, an
optimal input impedance at dual-band is obtained.

The layout and ultimate dimensions of the decoupled
monopole MIMO antenna are depicted in Fig. 7(a). With
respect to the layout of the decoupled monopole MIMO an-
tenna, it is worth mentioning that a post-tuning process of the
circuit parameters for the CRLH-TL is needed, due to the in-
ductor of vias ($\varphi = 0.5$ mm), the electrical length of pads (0.6
mm $\times$ 0.4 mm) and the capacitive coupling effect between
adjacent pads are not taken into account in circuit simulation.
The final circuit parameters of the CRLH-TL are $L_4 = 3.3$
H, $C_1 = 1.1$ pF, and $\theta_k = 70.68^\circ$, respectively. Fig. 8(a)
shows the simulated results comparison of the two-port S-
parameters between the reference and decoupled monopole
MIMO antenna. Compared with the reference antenna, the
mutual coupling coefficient $S_{21}$ is improved by 20 dB at two
center resonance frequencies (3.5 GHz and 4.9 GHz). Fur-
thermore, the isolation $S_{22}$ of two ports drops to below -20
dB at both two frequency bands (3.39 GHz - 3.59 GHz, 4.81
GHz - 5.01 GHz). Meanwhile, the reflection coefficient $S_{11}$
is better than -10 dB and the CCD of two antenna elements
is only 0.11$\lambda_2$, showing good performance of the proposed
dual-band decoupling scheme.

3.3 Performances of the Decoupled MIMO Antenna Array

3.3.1 S-Parameters

To further validate the decoupling performance of the pro-
duced dual-band decoupling strategy used in the closely
spaced two-element dual-band MIMO antenna, a prototype
of the decoupled monopole array is fabricated and presented

Fig. 7 (a) Layout and (b) prototype of the decoupled monopole MIMO antenna with $w_1 = 1.2$, $w_2 = 2.2$, $w_3 = 1.3$, $w_m = 0.7$, $l_1 = 2.45$, $l_2 = 19.6$, $l_3 = 18.1$, $l_4 = 12.4$, $l_m1 = 1.7$, $l_m2 = 11.8$, $l_m3 = 2.05$ (unit: mm).

Fig. 8 (a) Simulated S-parameters for the reference and decoupled monopole MIMO antenna. (b) Simulated and measured S-parameters of the decoupled monopole MIMO antenna.

Fig. 9 (a) Simulated and measured envelope correlation coefficient of the decoupled monopole MIMO antenna. (b) Diversity gain for the decoupled and reference monopole MIMO antenna.
in Fig. 7(b). Herein, the Murata surface mount technology (SMT) chip components with a package size of 0402 (1 mm × 0.5 mm) are adopted to make up the LH part of the CRLH-TL, and the Agilent N5244A vector network analyzer is employed to test the two-port S-parameter of the decoupled monopole array. As depicted in Fig. 8(b), the measured results are well consistent with the simulation ones. The discrepancy, specifically the slight frequency shifting in the upper band, can originate from the fabrication tolerance and the variation of the dielectric constant of the FR4 substrate.

3.3.2 Envelope Correlation Coefficient and Diversity Gain

The envelope correlation coefficient (ECC) is a crucial parameter to evaluate channel isolation for a MIMO antenna, which can be derived from either the radiation patterns or S-parameters. In this work, for simplicity, ECC calculated by the S-parameters is chosen[5]. According to the definition of ECC for two-port S-parameters, it is

\[
\text{ECC}(\rho_e) = \frac{|S_{11}^* S_{12} + S_{21}^* S_{22}|^2}{1 - |S_{11}|^2 - |S_{21}|^2 - |S_{22}|^2}.
\] (10)

As illustrated in Fig. 9(a), the simulated and measured ECCs of the decoupled monopole MIMO antenna are all below 0.05 at both interested frequency bands, which are much lower than the criterion for practice to use (ECC < 0.5)[5]. It means that pretty good isolation is achieved by employing the proposed dual-band decoupling strategy.

For a well-decoupled MIMO system, the diversity gain (DG) can be utilized to estimate the diversity performance for weakly correlated MIMO antenna elements. Supposing the well-matched two-element MIMO antenna is lossless and in a uniform/isotropic random field case (the receiving antenna is under the same condition), DG can be inferred by the ECC calculated above as[25], [30]:

\[
DG = 10\sqrt{1 - \text{ECC}(\rho_e)}.
\] (11)

As depicted in Fig. 9(b), the performance of the DG for the well-decoupled MIMO antenna array approaches 10 dB at both interested frequency bands, which is evidently better than the reference one. At the same time, the DG of the decoupled MIMO antenna is inferior to that of the reference antenna ranging from 4.1 GHz to 4.5 GHz, this is because the isolation of the decoupled MIMO antenna is less effective to the reference one, despite the reflection coefficient of the decoupled MIMO antenna being lower than -10 dB.

3.3.3 TRAC and Capacity Loss

It is well known that the traditional S-parameter matrix cannot precisely describe the bandwidth for a MIMO antenna system, however, the total active reflection coefficient (TARC) takes the influence of the mutual coupling and incident wave phase into account, which is a better choice to characterize the performance of the whole MIMO antenna array[9], [25], [30]. The TARC can be calculated by

\[
\Gamma_a = \sqrt{\left(\left|S_{11} + S_{12} e^{i\theta}\right|^2 + \left|S_{21} + S_{22} e^{i\theta}\right|^2\right)}/\sqrt{2},
\] (12)

where \(\theta\) is the excitation phase angle.

Under the high signal-to-noise ratio circumstances, the channel capacity loss (CL) induced by the correlation matrix can be obtained as

\[
C_L = -\log_2 \det(\psi^R),
\] (13)

where \(\psi^R\) is the correlation matrix for a two-element MIMO antenna system, and the elements belonging to the correlation matrix can be inferred from

\[
\psi_{ij} = 1 - \left|S_{ij}\right|^2, i \neq j = 1, 2.
\] (14a)
As illustrated in Fig. 10, the tendency of the TRAC for the decoupled MIMO antenna and the reference antenna is similar to its S-parameters respectively. In the meantime, the CL of the decoupled MIMO antenna array is better than the referenced one at both two interested frequency bands.

3.3.4 Radiation Patterns and Peak Gain

Fig. 11 depicts the simulated and measured normalized far-field radiation patterns of the decoupled monopole MIMO antenna at 3.5 GHz and 4.9 GHz in the $xoz$ plane and $yoz$ plane, respectively. During the testing process, we keep one port (port1) of the MIMO antenna array activated while the other one (port 2) is terminated by a 50-ohm matching load.

The radiation patterns of the decoupled monopole MIMO antenna are quasi-omnidirectional and the measured peak gain of the decoupled monopole MIMO antenna are basically same as the simulated ones at both frequency bands, as shown in Fig. 12.

3.3.5 Decoupling Performance Comparison

In order to show the decoupling performance of the proposed dual-band decoupling strategy, a comparison list with different dual-band decoupling methods is listed in Table 1. Compared with the EBG in [9], TLDBN in [21], Metal strips in [23], decoupling circuits in [24], and resonant structure in [25], the proposed dual-band MIMO antenna shows better decoupling performance in the achieved BW ($S_{11}$<$-10$ dB, $S_{12}$&$S_{21}$ < -20 dB) and CCD between the antenna elements at both interested bands.

A 5G (sub-6G) dual-band monopole MIMO antenna is designed, decoupled, and fabricated, the measured results agree well with the simulation ones. The decoupled antenna prototype achieves a dual-band of 3.40 GHz–3.59 GHz and 4.79 GHz–4.99 GHz, moreover, the center-to-center space between antenna elements is only 0.11λ at 3.5 GHz, showing good decoupling performance for closely spaced dual-band MIMO antenna. Besides, the proposed dual-band decoupling method is based on the two-port S-parameters matrix, which is antenna independent and can be expanded to decouple other dual-band two-element MIMO antenna.

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