Symmetrical double-loop H-field probe with floating shield for improving sensitivity and electric field suppression

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Abstract
A magnetic near-field probe (H-field probe) with dual output has been developed using a four-layer printed circuit board (PCB) technique. To achieve the improvement of detection sensitivity, the symmetrical double-loop is adopted to double the detection signal while suppressing the electric field coupling, which is also illustrated in a circuit model. The calibration factor (defined as the ratio of the external magnetic field to the induced response voltage of the probe) of the proposed H-field probe reaches 37.74 [dB (A/m)/V] at 0.5 GHz with the loop area of 0.8 mm², resulting in a better sensitivity than the single-loop probe. For the suppression of unwanted electric field a floating shield is added to the bottom of the probe, leading to an excellent common electric field suppression ratio of 29 dB in the frequency range up to 10 GHz. The probe is also characterised by parameters such as high differential electric field suppression, spatial resolution and probe influence on the device under test.

1 INTRODUCTION
With the development of miniaturisation, high operation speed, high power integrity and high density of devices, electromagnetic interference (EMI) and signal integrity issues on digital printed circuit boards (PCBs) have become increasingly complicated. The recognition of interference path and ground loop, and the location of interference source have become the main challenges in improving the electromagnetic compatibility of compact electronic products [1–3]. A magnetic near-field (H-field) probe acting as a useful near-field diagnostic tool is now widely used for radiation emission measurement [4,5], EMI source localisation [6,7], time-domain voltage and current measurement [8–10], and magnetic field measurement [11–14].

A great deal of effort has been made to improve the performance of the magnetic near-field probe since a simple loop probe antenna was proposed for measuring the magnetic near-field early in 1964 [12]. For example, a popular shielded-loop probe has been designed for shielding electric field interference [15]. In order to improve the spatial resolution of a shielded-loop probe, a miniaturised H-field probe has been designed using a thin-film process [16,17]. Several wide-frequency-band probes with high electric field suppression ratio have been designed based on a flexible low-temperature co-fired ceramics (LTCC) process [18]. In 2016, using the PCB technique, a simple miniature wide-frequency-band shielded-loop probe with an optimised transmission line structure was designed [6], and the impedance matching transmission structure improved the transmission efficiency and sensitivity of the detection signal [6]. In order to suppress differential E-field coupling, an H-field probe with floating shield was also proposed [19] in 2018, while a larger area unshielded-loop probe is also proposed to improve sensitivity [9]. In 2019, by using a two-port vector network analyser, an efficient near-field scanning system was proposed [20] to simultaneously measure electric and magnetic fields with a dual probe. Meanwhile, with the help of
a two-port vector network analyser, a near-field measurement system with differential symmetrical shielded-loop probe was proposed also based on the PCB process, and a common electric field suppression was reached at 40 dB in the frequency ranging from 0.05 to 11 GHz [21]. Generally, it is interesting to design a magnetic field probe with high sensitivity and excellent electric field rejection for the measurement application with the coexistence of electric and magnetic fields. In 2020, to obtain high electric field suppression the ultrawideband differential magnetic near-field probe was designed [22] with shielding vias to suppress the common-mode electric field. However, the design [22] of the shielding vias connecting the top and bottom ground planes in the bottom of the detection part could increase the electric field coupling, especially the differential electric field coupling. The vias also lead to a larger distance between the detecting loop and the device under test.

In order to characterise the performance of the probe, there are several key parameters for characterisation, such as frequency response, cross-polarisation, calibration factor, common electric field suppression, differential electric field suppression and spatial resolution. The authors focus on two parameters, which are calibration factor and electric field suppression. Usually, a large loop is good for high sensitivity while reducing spatial resolution. The sensitivity is related to the calibration factor, which is defined as the $H/V_0$, where $H$ is the magnetic field to measure and $V_0$ is the output voltage of the probe. The calibration factor at low frequency ($f$) is high because of the weak coupling at a low angular frequency ($\omega = 2\pi f$). The calibration factor of a typical shielded-loop probe with a 0.389 mm$^2$ loop is 47.5 dB at 2 GHz, and about 38.27 dB at 10 GHz [6]. The difference between the calibration factors of 10 and 2 GHz is 9.23 dB. In addition, the electric field suppression of the H-field probe is related to the design of the probing structure. The electric field suppression ratio is about 25 dB in frequencies below 10 GHz [6] for a typical shielded-loop probe made using the PCB process, while the electric field suppression ratio of the shielded-loop probe made using the LTCC process with added parallel C-shaped strips can reach 30–40 dB [18]. For the PCB fabrication technique, the thinnest diameter of wire is about 100 $\mu$m, which is roughly comparable with the LTCC fabrication technique. In addition, LTCC fabrication shows superiority in layer resource and flexibility in layer design. However, it is usually much more expensive than PCB fabrication.

A symmetrical double-loop H-field probe with a floating shield is proposed for sensitivity and electric field suppression improvement. The probe is easily achieved, fabricated with a four-layer PCB, and of low cost. The structure and detailed circuit model of the proposed probe are described in Section 2. In Section 3, the authors measure and validate the characteristics of the proposed probe by using a microstrip line providing a quasi-TEM field. Meanwhile, the characteristics of the proposed probe are compared with those of the reported probes, showing that the proposed probe has excellent sensitivity and electric field suppression. The conclusions are drawn in Section 4.

### 2 Principles of the H-Field Probe

#### 2.1 Popular single-shielded-loop probe

A popular single-shielded-loop probe [5,6] manufactured by a four-layer PCB technique is shown in Figure 1a. The probe consists of two shielded-ground plates on the top and bottom layers, a thin inner conductor and a via. In order to capture the magnetic field, two loop apertures are dug out from the shielded-ground plates. The inner conductor twines the aperture with area $S$ and then connects to the shielded-ground by the via.

To describe the detection principle, the self-inductances on each side of the shielded-ground plate are defined as $L_{g1}$ and $L_{g2}$ (shown in Figure 1a), the self-inductance of the inner conductor as $L_o$, and the mutual inductance between $L_{g2}$ and $L_o$ as $M_{g2o}$. The equivalent circuit diagram is shown in Figure 1b.

For the magnetic field probe, the magnetic field intended for detection is $H \cdot \sin \omega t$ (abbreviated as $\mathbf{H}$) in the horizontal direction and the unwanted electric field is $\mathbf{E}$, mainly in the vertical direction. A signal analyser with internal resistance $Z_L$ is used to measure the differential-mode voltage $U_0$. According to the electromagnetic induction law, the voltage induced by the magnetic field $\mathbf{H}$ can be expressed as $U_m = j\omega \mu H \cdot S$, where $\omega$ is the angular frequency of the activating field, $\mu$ the permeability of the loop antenna, $\mathbf{H}$ the magnetic field normal to the plane of the loop, $N$ (here $N = 1$) the number of loop, and $S$ the area of the loop. $U_m$ is embedded in the inner conductor circuit in the form of a mutual inductance voltage source (shown in Figure 1b). The output voltage $U_0$ can be written as $j\omega \mu \mathbf{H} \cdot S \mathbf{A}$, where $A$ is a system constant and determined by $L_o, L_{g2}, Z_L, C_s$ and the transmission structure of the probe.

The influence of the electric field coupling can be equivalent to a current source $I_e$. The nonsymmetric structure of the magnetic field probe can lead [6,18,23] to the unbalanced currents $I_{13}$ and $I_{23}$ (shown in Figure 1a), which causes the conversation between the common-mode noise to the measured differential mode signal. The conversation of the differential mode signal caused by the unwanted electric field coupling is then mixed with the signal $U_m$ and results in a low electric field suppression ratio especially at high frequency.

With ideal signal transmission, the contribution of electric field coupling to the output voltage $U_o$ is equivalent to a current source $I_{eq}$ and then the equivalent circuit diagram can be simplified as Figure 1c. The following formula can be obtained.

$$U_o = \frac{j\omega \mu H S}{j\omega L_{sa} + Z_L} Z_L' - I_{eq} Z_L', \quad (1)$$

where $L_{sa}$ is the sum of all inductors in the loop and equal to $L_e + L_{g2}$, and $Z_L'$ is the total load contributed by $Z_L$ and $C_s$ in parallel.
When the angular frequency is low enough \((Z_0L \gg j\omega L_{sa})\), the voltage \(U_o\) in Equation (1) can be approximated to be

\[
U_o = -j\omega \mu H_S - I_{eq}Z_0L_{sa}, \tag{2}
\]

Equation (2) illustrates that at low frequencies the voltage \(U_o\) is basically proportional to the frequency. In other words, the response curve of the probe at low frequency increases linearly with the frequency.

When the angular frequency is high \((Z'0L \ll j\omega L_{sa})\), the output voltage \(U_o\) can be approximated to be

\[
U_o = -j\omega \mu H_S \frac{Z'}{L_{sa}} - I_{eq}Z'_0L_{sa}, \tag{3}
\]

Equation (3) indicates that at high frequencies the voltage \(U_o\) is independent of the angular frequency and the frequency response curve of the probe stays constant with frequency. Equations (2) and (3) can apply to a general magnetic near-field probe with detecting loop. The second term \(I_{eq}Z'_0L_{sa}\) in Equations (2) and (3) illustrates that the output of the probe is contributed partly by the unwanted electric field coupling.

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2.2 Proposed symmetrical double-loop probe

In order to improve the sensitivity and electric field suppression, a differential symmetrical double-loop H-field probe is proposed, as shown in Figure 2. In the probe the coax through-hole is adopted to achieve an impedance match and then the transmission performance [6]. The width of the inner conductor and thickness of the stack are also designed to ensure the transmission impedance of 50 \(\Omega\). The centrally symmetrical loops 1 and 2 are placed back-to-back to form a double-loop structure. Loop 1 is placed in the third layer and connected to the bottom layer through a buried hole, and similarly loop 2 is placed in the second layer and connected to the top layer through another buried hole. Finally, a floating shield is adopted for electric field suppression (shown in Figure 2).

Since the detector is symmetrical, thin enough (about 0.5 mm) and wrapped by the floating shield, the circuit parameters \((L_{sa}, C_{sa}, L_{g1}\) and \(M_{gs2}\)) of each loop are considered to be equal. Meanwhile, at a low-enough frequency with electrical small-size approximation, the electric field coupling (either differential mode or common mode) is uniformly distributed along the floating shield and acts on the detection structure. In
this symmetric model, both loops couple with the electric field simultaneously, and the voltages of each port induced by electric field coupling are assumed to be the same. Similarly to Figure 1c, the simplified circuit model of the proposed probe with a floating shield is shown in Figure 3. The voltage induced in loops 1 and 2 can be expressed as $U_{o2} = -j\omega \mu H S$ and $U_{m1} = j\omega \mu H S$. The equivalent influence $I_{eq}$ induced by electric field coupling on both loops can be eliminated by differential operation $(U_{o2} - U_{o1})$ [21]. Therefore,

$$U_{o2} - U_{o1} = 2j\omega \mu H S A,$$  \hspace{1cm} (4)  

$$H = \frac{U_{o2} - U_{o1}}{2j\omega \mu S A},$$  \hspace{1cm} (5)

where

$$A = \frac{Z'_L}{j\omega (L_{sa} + M_s) + Z'_L}.$$  \hspace{1cm} (6)

The differential voltage between the two ports of the probe, $U_d = U_{o2} - U_{o1}$, is defined and then it can be calculated directly by a vector network analyser with embedded differential operation module. If the frequency is low enough ($Z'_L \gg j\omega (L_{sa} + M_s)$), the voltage $U_d$ can be simplified to

$$U_d = 2j\omega \mu H S,$$  \hspace{1cm} (7)

While the frequency is high ($Z'_L \ll j\omega (L_{sa} + M_s)$), the output voltage $U_d$ can be expressed as

$$U_d = 2j\mu H S \frac{Z'_L}{L_{sa} + M_s}.$$  \hspace{1cm} (8)

Equation (4) shows that the unwanted electric field coupling can be eliminated by the differential operation in the case of the perfect symmetry of both loops. For sensitivity, Equation (4) shows that the output differential voltage $U_d$ is twice as high as the voltage induced in a single loop (loop 1 or loop 2) in the completely symmetric circuit model. In other words, the frequency response of the double-loop can be improved by up to 6 dB in principle.

To evaluate the detection efficiency of a general magnetic field probe, $\gamma$ is defined as the output voltage over the loop area of the probe. Then the proposed symmetrical double-loop magnetic field probe can be compared to the published magnetic field probe with a single loop. For the proposed magnetic field with a double loop, $\gamma$ is approximated to be $2j\mu H Z'_L / (L_{sa} + M_s)$ at high frequency, while for the magnetic field with a single loop, $\gamma$ is approximated to $j\mu H Z'_L / L_{sa}$ by neglecting the electric field coupling. Note that the mutual inductance between the loops will cause deterioration of the sensitivity of the proposed symmetrical double-loop magnetic field probe.

3 | CHARACTERISTICS OF THE PROPOSED PROBE

3.1 | Frequency response and calibration factor

The system (Figure 4) for characterisation of the magnetic field probe has been widely used [4,5,19,20,24–26] for near-field
probe calibration, and it is mainly composed of a network analyser and a microstrip line. The microstrip line connects to a matched load and generates a quasi-TEM mode electromagnetic field for calibration. Recently, it was reported [26] for accurate probe characterisation using a microstrip line with air substrate, because the non-TEM field components are notable on the traditional dielectric substrate in high frequency. Such an air-substrate microstrip line is not easy to fabricate. Nevertheless, a RO4350B-substrate microstrip for probe characterisation at low frequencies is acceptable [4,5,19,20,24–26]. The magnetic field above the microstrip line can be calculated by [4,5,24].

$$H = \frac{U_3 b}{50\pi(d + 2b)}.$$  

where $U_3$ is the voltage injected from the microstrip line (voltage of port 3 in Figure 4), $b$ the distance between the trace and grounded plane of the microstrip line, and $d$ the distance between the trace and the centre of the detecting loop.

The differential output voltage $U_d = U_{o2} - U_{o1}$ can be calculated by

$$U_{o2} - U_{o1} = S_{23}U_3 - S_{13}U_3. \quad (10)$$

The frequency response of the proposed probe is defined as $S_d = S_{23} - S_{13}$, then the calibration factor $CF$ can be evaluated in dB as follow [5,20].

$$CF = \frac{H}{U_{o2} - U_{o1}}$$

$$= 20\log_{10}\left[\frac{b}{\pi d(d + 2b)}\right] - |S_d| - 34 \text{ dB} \cdot (\text{A/m})/\text{V}. \quad (11)$$

Figure 5 shows the frequency response of the probe and the calibration factor with the parameters $d = 1.7$ mm, $\theta = 0^\circ$.

To validate the measurement, the electromagnetic field simulation is performed with the model shown in Figure 6. In the model, it is designed with a full-wave simulation tool, and the port is set as the lumped port. It can be seen from Figure 5 that the simulation model can basically give the response trend of the probe in electromagnetic field detection. With the model, the effect of $d$ (the distance from the centre of the loop to the microstrip line) on the $CF$ can be also studied, as shown in Figure 7, where the parameter $d$ is increased from 0.9 to 2.4 mm with a fixed $b = 0.762$ mm and $CF$ changes slightly for $d$ from 1.7 to 2.1 mm. The circuit model has been verified by full-wave simulation tool and supports the probe design.

To show the characteristics of the proposed probe, a comparison of the sensitivity of various magnetic field probes in published papers is shown in Table 1. Since a larger loop would be good for the field detection and bad for the spatial resolution, a new parameter is defined as the product of the $CF$ and area of the loop for the comparison of probes given in Table 1. This
parameter indicates the detection efficiency of the probe with an equal loop area and is proportional to \(1/\gamma\). As can be seen, the probes proposed have good detection performance, that is 37.74 dB at 0.5 GHz, 26.71 dB at 2 GHz, 22.82 dB at 5 GHz, and 22.90 dB at 10 GHz in comparison to the other four single-loop probes in the table. This means that with the same loop area the proposed design of the probe can achieve higher sensitivity at low frequency. At high frequencies (10 GHz), the proposed magnetic field probe shows close performance in detection efficiency to the two probes in [21]. This can be explained by their close values of \(\gamma\). For the proposed magnetic field with a double loop, even though the differential operation of the proposed magnetic field with a double loop can double the detected signal, the detected signal would decrease at high frequency due to the mutual inductance \(M_s\) of the double-loop structure.

Compared with the double-loop probe in [22], the larger the loop, the lower the calibration factor; the smaller the loop, the higher the detection frequency. By adding the correction factor of \(1/\sqrt{2}\) for the definition of the response amplitude [22], the value of \(M_s\) of the probe proposed is basically the same as that in reference [22]. When the frequency is greater than 10 GHz, mutual inductance will reduce the detection sensitivity of the double-loop probe, which has no advantage over the single probe; at the same time, the asymmetry of the double-loop probe will also limit the working frequency band of the probe, which makes it difficult to design the dual-loop probe at higher frequency. Therefore, the method to realise the advantage of the double-loop design is to design the near-field probe with low frequency and large loop area. Compared with the design in [22], this probe design has more advantages in low-frequency detection.

## 3.2 | Cross-polarisation

Cross-polarisation of the probe can be reflected by the relationship between the response voltage \(U_d = U_{o2} - U_{o1}\) and \(\theta\), which is defined as the angle of the probe and trace (shown in Figure 4). When the loop plane is parallel to the microstrip line, the angle \(\theta\) is defined as 0°. As the probe rotates along the Z-axis (Figure 8), the response voltage of the probe would be \(U_d = 2j\omega y H S \cos(\theta)\). When the probe is placed at \(\theta = 90^\circ\), the magnetic field coupling is supposed to be zero (to a minimum value). Suppose that the electromagnetic field applied to the probing loop is uniform. Then in physics the response at \(\theta = 90^\circ\) would result from conversion (\(\gamma\)) by the common mode to differential mode signal. Such conversion is governed by the symmetry of both loops in the probe. While at \(\theta = 0^\circ\) the probe response is contributed not only by the common mode to the differential mode signal but also the magnetic field induction. The minimum-normalised response of the probe is defined as the ratio of \(U_d(\theta)/U_d(\theta = 90^\circ)\) (Figure 8). Figure 8 shows the minimum-normalised cross-polarisation of the probe without and with the floating shield, respectively. Here, the step of \(\theta\) is set at 5°. As can be seen from Figure 8, with the increase in frequency, the minimum-normalised response of the probe increases. This means that the response of the probe is much more distinct for \(\theta = 0^\circ\) than \(\theta = 90^\circ\). To the best of the authors’ knowledge, such distinction with frequency is due to the symmetry deterioration of both loops in the probe at high frequency.

Figure 8 also shows that the cross-polarisation of both probes with and without a floating shield basically satisfies the cosine changing trend. The maximum of minimum-normalised received voltage of the proposed probe without a floating shield are 12.76 dB at 10 GHz, 16.42 dB at 5 GHz and 18.16 dB at
TABLE 1 Performance comparison among magnetic probes

| Probes          | [18]     | [6]    | Ref. Probe in [21] | Proposed in [21] | Proposed in [22] | Proposed Probe |
|-----------------|----------|--------|--------------------|-------------------|-------------------|----------------|
| Manufacture technique | LTCC | PCB    | PCB                | PCB               | PCB               | PCB            |
| Area of loop    | 0.04 mm\(^2\) (Single-loop) | 0.389 mm\(^2\) (Single-loop) | 0.360 mm\(^2\) (Single-loop) | 0.360 mm\(^2\) (Single-loop) | 0.180 mm\(^2\) (Double-loop) | 0.800 mm\(^2\) (Double-loop) |
| Operating frequency | 0.01 ~ 20 GHz | 9 kHz ~ 20 GHz | 0.05 ~ 20 GHz | 0.05 ~ 20 GHz | 0.1 ~ 14 GHz | 0.05 ~ 10 GHz |
| $CF [\text{dB}(\text{A/m})/\text{V}]$ | | | | | | |
| At 0.5 GHz      | 85.92 dB | 62.50 dB | 50.11 dB | 49.24 dB | 53.36 dB | 37.74 dB |
| At 2 GHz        | 71.72 dB | 47.50 dB | 38.39 dB | 37.37 dB | 42.27 dB | 26.71 dB |
| At 5 GHz        | 64.43 dB | 41.30 dB | 33.13 dB | 31.54 dB | 37.75 dB | 22.82 dB |
| At 10 GHz       | 59.65 dB | 38.27 dB | 29.06 dB | 28.91 dB | 36.81 dB | 22.90 dB |
| $CF \cdot \text{Area} [\text{dB}(\text{A/m})/\text{V} \cdot \text{mm}^2]$ | | | | | | |
| At 0.5 GHz      | 57.96 dB | 54.30 dB | 41.24 dB | 40.30 dB | 38.47 dB | 35.80 dB |
| At 2 GHz        | 43.76 dB | 39.30 dB | 29.52 dB | 28.50 dB | 27.38 dB | 24.77 dB |
| At 5 GHz        | 36.47 dB | 33.10 dB | 24.26 dB | 22.67 dB | 22.86 dB | 20.88 dB |
| At 10 GHz       | 31.69 dB | 30.07 dB | 20.19 dB | 20.04 dB | 21.92 dB | 20.96 dB |

Abbreviations: CF, Calibration factor; LTCC, low-temperature co-fired ceramics; PCB, printed circuit boards.

![Figure 8](image.png)

**FIGURE 8** Cross-polarisation suppression of the probe with and without a floating shield at $d = 1.7$ mm

1 GHz, while the maximums of minimum-normalised received voltage of the proposed probe with a floating shield are 31.54 dB at 10 GHz, 36.14 dB at 5 GHz and 38.23 dB at 1 GHz. The cross-polarisation of the probe with a floating shield is much better than that of the probe without a floating shield. To discover the cause of the difference in cross-polarisation between the two kinds of probes, the response voltage $U_d$ at 5 GHz of the probe without a floating shield is recorded in Figure 9a, when the probe is placed at $\theta = 90^\circ$ and moved from $Y = -4$ mm to $Y = +4$ mm. The shape of the recorded response voltage (Figure 9a) is basically consistent with the magnetic field distribution of the microstrip line, indicating the possible magnetic induction lines in this situation. Such a magnetic induction by the hidden loop (formed by the buried vias) is sketched in Figure 9b. Since the coupling in Figure 9b can be suppressed by adding a floating shield for prevention of the magnetic induction lines from going through the buried holes, the probe with a floating shield can achieve a high cross-polarisation suppression.

### 3.3 Electric field suppression

The electric field suppression is used to describe the ability of the magnetic field probe to suppress uniform vertical electric field interference during a magnetic field measurement. In an ideal case of $\theta = 90^\circ$, the response voltage of the magnetic field probe is expected to be 0, while in the case of $\theta = 0^\circ$, the response voltage of the magnetic field probe should reach a maximum contributed completely by the magnetic field coupling. In practice, the electric field suppression ratio of a magnetic field probe is defined as the ratio of the response voltage measured at $\theta = 0^\circ$ to the one measured at $\theta = 90^\circ$ [18]. The electric field suppression ratio of a typical magnetic field probe with a shield-loop is generally 23 dB [6,21]. Figure 10 shows the electric field suppression ratio of the proposed probe with floating shield. As can be seen from Figure 10, the electric field suppression ratio of the proposed probe can reach 29.52 dB in the measurement of frequencies from 0.2 to 10 GHz, and is higher than 35 dB in the frequency below 7.25 GHz. Although the floating shield can increase the electric field suppression, the proposed probe has a moderate electric field suppression in the studied frequency range. The differential symmetrical magnetic field probe proposed in [21] has the highest electric field suppression of 40 dB. The traditional shielded-loop magnetic field probe in [6,9,21] has an electric field suppression ratio less than 25 dB, and the probe made using the LTCC process in [18] has 30 dB electric field...
SUPPRESSION AND DIFFERENTIAL ELECTRIC FIELD suppression. Nevertheless, the floating shield can be a choice for the further improvement of electric field suppression in [6,9,18, 21].

3.4 | Differential electric field suppression

Different from the electric field suppression caused by uniform electric field interference, the differential electric field suppression is caused by a non-uniform electric field. The differential electric field suppression is a performance indicator for near-field scanning [19]. When the H-field probe moves to the edge of the measured trace (show in Figure 11), the non-uniform E-field is imposed on the detector of the probe and causes a differential mode current and then affects the output of the probe through differential current.

The differential electric field suppression is defined as the ratio of the response voltage measured at θ = 0° to the one measured at θ = 90°. When the probe is placed at θ = 90° and moved across the microstrip line from Y = 3800 μm to Y = 0° with the lift off of 800 μm, the response voltage of the probe is plotted as a function of the position at 10 GHz in Figure 12. Similarly, the curve measured at θ = 0° is recorded, and also shown in Figure 12. It can be seen from Figure 12 that the differential electric field suppression is 10 dB at 10 GHz. Figure 13 shows the differential electric field suppression in the frequency from 0.2 to 10 GHz. Usually, when the differential electric field suppression ratio is lower than 0 dB, the near-field scanning result of the magnetic field would be misleading due to the possibly existing electric field interference. Then the source localisation according to the near-field mapping in this situation could be wrong.

3.5 | The influence on the microstrip line

When the probe is placed close to the device under test (DUT), the influence of the probe on the DUT should be considered. The transmission (the amplitude and phase of S21) variation of the microstrip line is used to characterise the effect of the approaching probe on the microstrip line (as a standard DUT) [18]. Figure 14 shows the amplitude and phase variations of S21, when the distance between the probe and the microstrip line is set at 800 and 500 μm, respectively. It can be seen from Figure 14 that the influence of the approaching probe on the microstrip line is negligible (the amplitude difference 0.5 dB, and the phase difference 0.03°) in the measured distance. This indicates that the attached floating shield would not cause a significant effect on a DUT even at frequencies up to 10 GHz, if a proper working distance is set.

3.6 | Spatial resolution

To characterise the spatial resolution of the proposed probe, a microstrip line with a defined width is adopted as a
standardised sample, as shown in Figure 4. When the probe moves across the microstrip line (W = 1.55 mm) from Y = −3800 μm to Y = 3800 μm at θ = 0°, the magnetic field profile measured by the probe is recorded. The spatial resolution is defined as the distance from the peak position of the response voltage maximum to the position of the −6 dB decrease level [5,18,21]. Figure 15 shows the spatial resolution of the proposed probe, respectively, at 1, 5 and 10 GHz, when the probe is kept above the microstrip line at a height of 800 μm. Figure 15 shows that the spatial resolution of the proposed probe is about 1.3 mm in this case.

4 | CONCLUSIONS

To conclude, a symmetrical double-loop H-field probe with a floating shield is proposed. By using the double-loop design, the detection sensitivity can be improved by about 6 dB especially at low frequency, while little improvement is achieved due to the mutual inductance between the two differential loops. For electric field suppression, it can be improved by the differential operation, and meanwhile, the floating shield is attached to the bottom for further coupling suppression of the unwanted magnetic field passing through the buried hole. The total electric field suppression is greater than 29 dB at frequencies from 0.05 to 10 GHz. Other characterisations are also performed on the proposed magnetic field probe, such as high differential electric field suppression at about 10 dB, spatial resolution of 1.3 mm, and negligible influence on DUT. Generally speaking, the double-loop and floating shield design in the proposed probe would be constructive to improve the sensitivity and electric
field suppression of a magnetic field probe. Note that in practice the copper tape floating shield of the probe can be implemented as edge plating in the PCB fabrication technique.

To the best of the authors’ knowledge, there are usually two kinds of measurement purposes with a probe. One is to locate an EMI source, and the other is to measure the external magnetic field. For the former one, the measurement accuracy of the probe in quantity is not an important issue, since the hot spot and the profile of an EMI source could be enough for localisation. In this context, the spatial resolution and the electric field suppression could be of more concern, especially for a highly integrated and complex sample. For the latter one, the accuracy of the measurement of an external magnetic field depends on the error of the CF (calibration factor) curve. In this case, careful calibration is needed, such as an air-substrate microstrip line, a laser rangefinder for distance detection, and an accurate positioner to locate the probe to the centre above the microstrip line.

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CONFLICT OF INTEREST STATEMENT
The authors declared that they have no conflicts of interest to this work. They declare that they do not have any commercial or associative interest that represents a conflict of interest in connection with the work submitted.

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