行政院國家科學委員會專題研究計畫成果報告

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★ K-頻段無線收發關鍵元組件之研究(3/3) ★

★ **子計畫一：K-頻段雙頻印刷天線(3/3)** ★

★ Key Devices and Components for K-band Wireless Transceiver (3/3) ★

★ **Sub-Project 1: A K-Band Dual-Frequency Printed Antenna (3/3)** ★

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計畫類別： □個別型計畫 ■整合型計畫

計畫編號： NSC 89-2219-E-002-039

執行期間： 89 年 8 月 1 日至 90 年 7 月 31 日

子計畫主持人： 許博文

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處理方式： □可立即對外提供參考

（請打■） ■一年後可對外提供參考

□兩年後可對外提供參考

（必要時，本會得展延發表時限）

執行單位：國立台灣大學電信工程學研究所

中 華 民 國 九十 年 十 月 三十一 日

Abstract

To match the common planar structure in all the sub-projects of this integrated research project, we study in this sub-project a dual-band printed antenna that could be used in both 21 to 23 GHz and 24 to 26 GHz frequencies. The proposed antenna is a coplanar waveguide fed slot-loop coupled microstrip antenna. Without using any parasitic elements, this antenna can have a bandwidth several times wider than any typical slot coupled microstrip antenna. Also, the operating frequencies of the slot-loop and the microstrip antenna can be adjusted to lie separately in the two desired frequency bands. The theoretical analysis is based on the method of moments using rooftop basis functions, while the experimental trial and error will be called upon in the actual design to compensate for the discrepancy between theory and experiment in K-band. The method of analysis and the experimental setup had been established in the first year, the detailed design had been processed in the second year, and a prototype antenna that meets the design specifications has been completed in this (third) year. This report summarizes the final year's research results that by using the analysis method and the experimental setup developed in the previous years, we have tuned the antenna design in the K-band to a satisfactory result.

Keywords: Coplanar Waveguide, Slot Loop, Microstrip Antenna.

摘　要

為配合整合型計畫各子計畫之平面結構，本子計畫研究供21　23GHz及24　26GHz兩頻段均能使用之平面型印刷天線。採用之天線為共面波導饋入之環狀槽孔耦合微帶天線。此天線，在不使用寄生元件時，即可比一般直槽耦合微帶天線多數倍的適用頻寬。經由適當之設計，環狀槽孔及微帶天線之操作頻率可分別落於需要的兩頻段內而達到雙頻操作之目的。本研究之理論分析採用以屋頂形函數為基底之動差法，而實驗試錯法被用來做細部設計以彌補在K-頻段時理論之不足。計畫第一年以建立理論模型及實驗平台為主，第二年進行細部設計，第三年則完成符合規格之雛型成品。本報告為本計畫第三年之研究成果報告，利用前二年所建立之理論模型及實驗平台，在K-頻段加以細部微調，獲得相當不錯的設計結果。

關鍵詞：共面波導，環狀槽孔，微帶天線。

1.Introduction

In many advanced communication systems, such as the global positioning system (GPS), vehicular communication system, …and etc., require low profile antennas capable of dual frequency operation with sufficient broad bandwidth [1]-[3]. A dual-band coplanar waveguide (CPW)-fed slot-loop coupled microstrip antenna designed in this study will suffice these requirements.

The antenna configuration is shown in Fig. 1. A rectangular patch is placed on one side of the substrate, while a rectangular slot-loop fed by a CPW is arranged opposite to the patch in the ground plane on the other side of the substrate. The rectangular patch is placed centrally about the rectangular slot-loop. Because of its uniplanar configuration, the CPW facilitates parallel and series connection of both passive and active components. As a result, CPW lines are very suitable and competitive for the design of active integrated antennas. Moreover, CPW can yield advantageous propagation characteristics, i.e. small dispersion and low conductor and radiation losses up to millimeter-wave frequencies, under proper design [4]. According to the radiation mechanism, the resonant length of the patch and the total length of the rectangular slot-loop determine the resonant frequencies of the antenna, respectively [5]-[6]. By comparison with the dual-band microstrip antennas introduced in [2], the new technique presented in this paper provides an alternative method to design a dual-band microstrip antenna, which not only includes advantages of the CPW-fed microstrip antenna, but also make the fabricating procedure easier and simpler.

In the following, the radiation mechanism of the patch and the slot-loop is introduced to estimate the resonant frequencies of the antenna. And a strict theory based on the full-wave analysis is used to analyze the performance of the CPW-fed dual-band microstrip antenna. For design purpose, a parameter study for variations of the resonant frequency with the size of the structure is presented. To verify our method, we fabricate several dual-band CPW-fed slot-loop coupled microstrip antennas, then measure and discuss the differences between numerical calculations and experimental measurements. After a slight adjustment of the test piece, a dual-band microstrip antenna that has both the broad impedance bandwidth and the dual-band operation feature in the K-band is obtained.

2. Radiation Mechanism

To calculate the resonant frequency of the patch antenna, the simple lossless-cavity model is used [5]. Assuming no variations of the electric field along the width and the thickness of the microstrip structure, the electric field configuration of the radiator can be represented as shown in Fig. 2. The fields vary along the patch length, which is about half a wavelength. Radiation may be ascribed mostly to the fringing fields at the open-circuited edges of the patch. Based on a formula due to [7], the cavity resonates when the wavelength a_g is approximately half of the length of the patch. That is

(1)

where c is the speed of light, $f_{r,p}$ is the resonant frequency of the patch antenna, $L_{p,eff}$ is the effective length including fringing effect, and the other parameters are shown in Fig. 1. In general, the measured resonant frequency of the patch antenna is slightly lower than the computed one by using the equation (1) [5].

In the case of the slot-loop antenna, the energy of the electromagnetic field is considered to be mainly localized in the slot. The slot will become resonant when the average total length of the slot is approximate to a multiple of the wavelength. That is

(2)

where $f_{r,l}$ is the resonant frequency of the slot-loop antenna, N is an integer, and the other parameters are shown in Fig. 1. By using equations (1) and (2), the initial design parameters of the dual-band CPW-fed slot-loop coupled microstrip antenna could be determined immediately.

3. Theoretical Analysis

To simplify the analysis, we assume that the substrate and the ground plane extend to infinity in the x and y directions and that the conductors are perfectly conducting and of negligible thickness. The coordinates are chosen such that the ground plane coincides with the xy-plane, the CPW, the slot-loop, and the patch are all symmetric about the x-axis, and the y-axis is placed along the edge of the end of the CPW.

By invoking the equivalence principle and applying suitable boundary conditions, the magnetic field $\overline{H}^s$ in the slot aperture and the electric field $\overline{E}^p$ on the patch surface can be expressed as

(3)

(4)

(5)

(6)

where the superscripts a and b denote the fields or the sources just below and above the slot aperture surface, respectively, $\overline{J}^p$ and $\overline{M}^s$ are the electric and magnetic surface currents on the patch and in the slot aperture, respectively, $\overline{M}^{sb} = -\overline{M}^{sa}$, and the subscripts x and y represent the corresponding components of the fields. Each field in (3)-(6) is the field due to the specified current radiating in its corresponding region (region of half free-space or half-space with grounded dielectric slab). All the fields can be calculated from suitable Green's functions, whose spectral domain forms can be found from [8]-[10] or derived from the immittance approach [11]. To solve the unknown currents $\overline{J}^p$ and $\overline{M}^s$, a Galerkin moment method is used. After we obtain the unknown coefficients of $\overline{J}^p$ and $\overline{M}^s$, matrix pencil method is used to calculate the input reflection coefficient S_{11} [12]. For the slot magnetic current, only the longitudinal component is considered to simplify the analysis, which should be good enough for a narrow-slot assumption.

4. Results and Discussions

For the purpose of the design, it is important to discuss the relationship between resonant frequencies of the antenna and the sizes of the structure. The variations of the resonant frequencies with L_p and L_s are calculated and plotted in Figs. 3 and 4, respectively. From equations (1) and (2), the estimated resonant frequencies of the patch and the slot-loop are 12.76 GHz and 20.90 GHz, respectively. Comparing to the data presented in Figs. 3 and 4, it is found that the resonant frequencies of the antenna are determined mainly by the resonant length of the patch L_p and the length of the slot-loop L_s, respectively. We could suppose that the lowest resonant frequency is due to the TM_{10} resonant mode of the patch antenna, and the bandwidth defined for VSWR $\leq$ 2 is due to the resonances of the slot-loop and the TM_{20} resonant mode of the patch.

Then, we use parameters G_s = 2.5 mm, W_s = 0.25 mm, L_p = 6.5 mm, W_p = 6.0 mm, L_s = 4.6 mm, d_s = 3.8 mm, O_s = 1.0 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $\tan\delta$ = 0.0022 to verify our method. The experiment is carefully measured using HP8510B network analyzer with TRL (Through-Reflection-Line) calibration technique. The measured and calculated reflection coefficients S_{11} referenced to the end of the CPW are shown in Fig. 5. According to aforesaid statement, we could suppose that the resonant frequency, 11.6 GHz, of the experimental data is due to the TM_{10} resonant mode of the patch. And the bandwidth between 18.8 and 22.1 GHz is due to the resonance of the slot-loop and the TM_{20} resonant mode of the patch. The agreement between theoretical results and experimental data is better in the lower frequency range. The discrepancies between theory and experiment may be attributed to a number of sources such as the tolerances in the dimensions of the test piece, the finite size of the substrate, and the negligence of the conductor thickness. Also, in modeling the slot magnetic current, the transverse variation in the slot was assumed to be uniform disregarding the singular-edge behavior. As the operating frequency is increased, the mismatch between the theoretical result and the experimental data becomes more apparent. Moreover, according to equation (1), the unexpected resonant frequency, about 27 GHz, of experimental data is judged due to TM_{02} resonant mode of the patch.

Up to now, various microstrip antennas have been proposed to increase the impedance bandwidth or to be used for dual-band operation. But the combinations of these two objects are few. Based on the aforementioned procedure, we use G_s = 2.5 mm, W_s = 0.25 mm, L_p = 2.1 mm, W_p = 4.0 mm, L_s = 5.0 mm, d_s = 2.3 mm, O_s = 0.5 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $\tan\delta$ = 0.0022 to fabricate a dual-band microstrip, which has both the broad bandwidth and dual-band operation feature. The experimental data is plotted in Fig. 6, and the measured bandwidth of the dual-band CPW-fed slot-loop coupled microstrip antenna is wider than 21.46%.

5. Conclusion

In this paper, a simple technique to implement a dual-band CPW-fed slot-loop coupled microstrip antenna is presented. The theoretical calculation and the experimental measurement verify its validity. By comparison with other dual-band microstrip antennas introduced in [2], the new technique presented in this paper not only includes advantages of the CPW-fed microstrip antenna, but also makes the fabricating procedure easier and simpler. Moreover, the two resonant frequencies of the patch and the slot-loop could be designed closer or farther to obtain either the broad impedance bandwidth or the dual-band operation feature.

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Figures

(a)

(b) (c)

Fig. 1 Configuration of the CPW-fed dual-band microstrip antenna. (a) Perspective. (b) Top view. (c) Bottom view. ε_r is the dielectric constant of the substrate, and tan δ is the loss tangent of the substrate.

(a)

(b)

Fig. 2 Electric field configuration of the radiator. (a) Rectangular microstrip patch antenna. (b) Side view.

(a) (b)

Fig. 3 Variation of reflection coefficient with L_p. (a) Magnitude. (b) Phase. G_s = 2.5 mm, W_s = 0.25 mm, W_p = 6.0mm, L_s = 4.6 mm, d_s = 3.8 mm, O_s = 1.0 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $tan\,\delta$ = 0.0022.

(a) (b)

Fig. 4 Variation of reflection coefficient with L_s (a) Magnitude. (b) Phase. G_s = 2.5 mm, W_s = 0.25 mm, L_p = 6.5 mm, W_p = 6.0 mm, d_s = 3.8 mm, O_s = 1.0 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $tan\,\delta$ = 0.0022.

(a) (b)

Fig. 5 Reflection coefficient of a CPW-fed dual-band microstrip antenna. (a) Magnitude. (b) Phase. G_s = 2.5 mm, W_s = 0.25 mm, L_p = 6.5 mm, W_p = 6.0 mm, L_s = 4.6 mm, d_s = 3.8 mm, O_s = 1.0 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $tan\,\delta$ = 0.0022.

(a) (b)

Fig. 6 Reflection coefficient of a CPW-fed dual-band microstrip antenna. (a) Magnitude. (b) Phase. G_s = 2.5 mm, W_s = 0.25 mm, L_p = 2.1 mm, W_p = 3.5 mm, L_s = 5.0 mm, d_s = 2.3 mm, O_s = 0.5 mm, W_l = 0.5 mm, d = 1.58 mm, ε_r = 2.6, and $tan\,\delta$ = 0.0022.