Fully Metallic Compound Air-fed Array Antennas for 13 GHz Microwave Radio-link Applications

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***Abstract*- A novel structure of metallic compound air-fed array (CAFA) antennas, as a kind of improved Fabry-Perot resonator (FRP), is proposed in this paper. With stepped metallic ground plate, metallic grating cover and open-ended waveguide feeding, this kind of antenna exhibits features of low profile, low cost and easy integration into microwave out-door unit (ODU). The operating principle, design procedure, simulation analysis and an experimental prototype for application of 13 GHz microwave radio-link are presented. The measured results exhibit: 19.2 dBi peak gain, 3.85% common bandwidth for both 3 dB gain-drop and VSWR≤1.5:1, and good radiation characteristics.**

I. INTRODUCTION

The concept of CAFA antenna, which could be considered as an improved FPR antenna, has been proposed and studied [1] [2] recently. Compared to traditional FRP antenna, the most essential improvements of CAFA include bandwidth enlargement and/or aperture efficiency enhancement by adopting non-uniform base and/or cover to improve the phase distribution of aperture field [3]-[6]. Most of researches on CAFA so far were based on printed structure, so called as printed CAFA (P-CAFA). P-CAFA owns features of easy fabrication and integration with circuit, but also has some limitations in the out-door applications because of its insufficiency of strength and high cost of microwave dielectric material. In this paper, a metallic CAFA (M-CAFA), which is based on full metallic structure, will be developed to extend its application to radio-link.

Some called metallic EBG resonator antennas, which have similar structure with M-CAFA, had also been studied. They were usually designed for special applications, such as generation of circularly polarized antennas [7] [8] or high performance feed of reflector antenna [9] [10]. But they also had more or less structural encumbrance with additional polarizer or feeding structure, which caused relative high profile and large aperture size and complex feeding. M-CAFA antennas proposed here will exhibit attractive features of compact size, design flexibility, potentiality of wide bandwidth and high aperture efficiency.

Microwave radio-link networks have become the lifeblood of the telecommunications industry, particularly with the rise of high traffic volumes of voice and data for 3G/4G system. One of the main objectives is to provide fast and cost-effective deployment of communications link. Antenna is one of the main components of radio-link system, and usually its size will determine the total size of system to a large extent. Most of the commercial microwave antennas are parabolic antennas, which usually own bulky structure and relative high manufacture cost. The scheme proposed in this paper will provide a very economical solution. Additionally, due to full metallic structure with low profile and quasi-planar configuration, this kind of antenna has high reliability and can be easily integrated or installed with out-door unit (ODU) or microwave system.

The antenna configuration, design principle and parameter analysis are presented with discussion in detail. A prototype for application of 13 GHz microwave radio-link is designed, fabricated, and measured for verification.

II. ANALYSIS AND DESIGN

A. *Configuration*

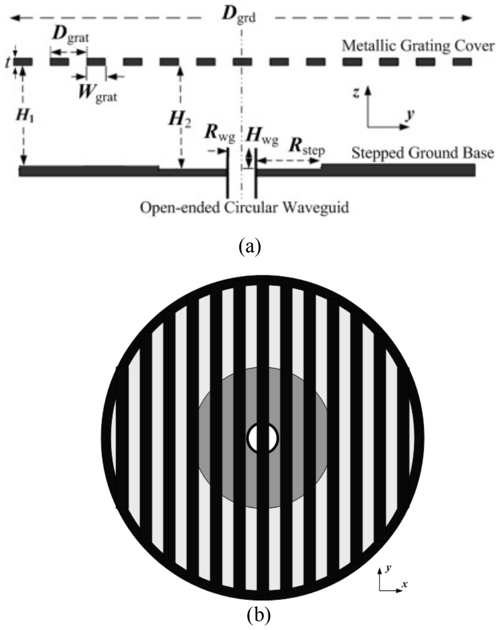


Fig.1. Configuration of proposed M-CAFA antenna. (a) Cutting view with main structural parameters definition. (b) Top view.

The proposed M-CAFA antenna is shown in Fig.1 with definition of most of geometry parameters. It consists of three main parts: a cover of partially reflective surface (PRS) consisting of metallic grating; a base of metal ground with one or more concentric steps; and an open-ended circular waveguide as feed protruded from the centre of base. The outline of antenna shown here is circular with diameter of D_{grd}, but it can be also designed into other shape according to required beam shape. The thin grating cover is designed with uniform size: period of grating D_{grat} and width of grating W_{grat}. But the ground base is designed with steps $H_{step}=H_2-H_1$, corresponding to different spacing H_1 and H_2 between cover and base, and inner radius of step R_{step}. The circular waveguide is designed to operate in dominant mode of TE_{11} and partly protruded from the base. The radius and insertion length of waveguide are donated as R_{wg} and H_{wg}. The required operating frequency is at f_0=13.0 GHz with bandwidth from 12.75 GHz to 13.25 GHz.

B. Initial Design for Traditional FPR Antenna

As well-known, for a traditional FPR antenna with uniform cover and base, if denote their reflection phase as $\phi_1(f)$ and $\phi_2(f)$ and the reflectivity as $r(f)$ and 1, respectively, then the resonant condition can be presented as:

$$\phi_1(f) + \phi_2(f) = \frac{4\pi H}{c} f - 2N\pi \quad (N=0,\ 1,\ 2\ ...) \qquad (1)$$

Where c is the light velocity, N is the order of resonant mode.

On the other side, its directivity gain D is determined by the reflectivity of cover as [11]:

$$D = [1+ r(f)] / [1- r(f)] \qquad (2)$$

So, the first step of design is to adopt properly the sizes of grating cover. By considering the operating frequency, realized gain according to (2) and engineering feasibility, the parameters of cover are optimally designed as: D_{grat}=10 mm, W_{grat}=4mm, with thickness t=0.5mm. The reflection magnitude and phase at the cover are shown in Fig.2, in which its reflection phase $\phi_1(f)$=(155~160)^o and magnitude from 0.96 to 0.97 corresponding to reflectivity $r(f)$=0.92~0.94, covering the frequency range of 12.0~14.0 GHz.

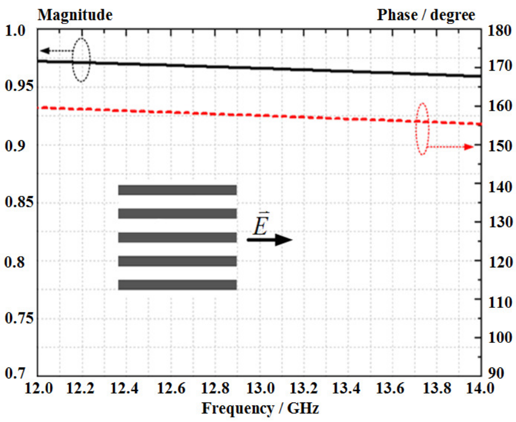


Fig. 2.Reflection coefficients of grating cover. (D_{grat}=10 mm, W_{grat}=4 mm)

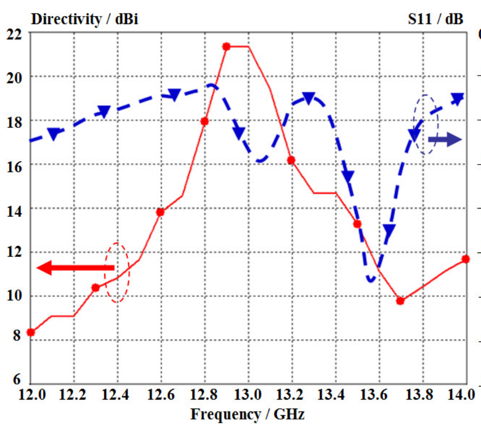
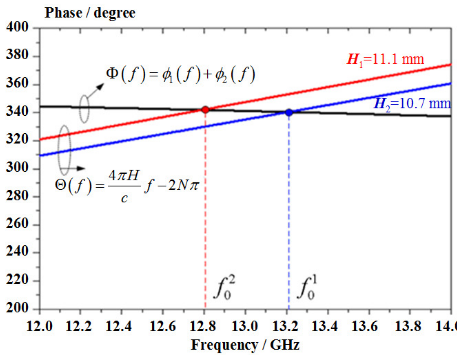


Fig.3. The initial performances of traditional FPR design.

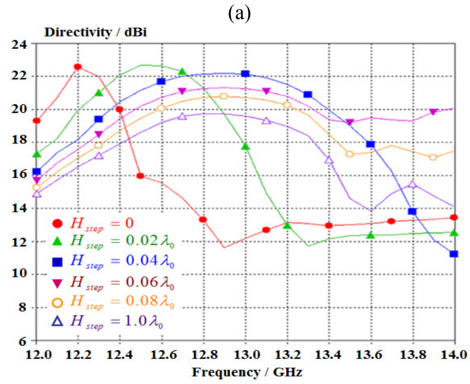
Set the original parameters as: aperture diameter D_{grd}=140 mm (6.0 λ_0), radius of waveguide R_{wg}=8.3mm (0.36 λ_0) and insertion length of waveguide H_{wg}=3.5mm (0.15 λ_0). In the case of ground base without step, the spacing should be calculated from (1) as H_2=H_1=11.0 mm (0.48 λ_0). By employing CST-2006 full-wave simulator, the initial performances of traditional FPR design are shown in Fig.3.

Although this design performs 21.7 dBi peak gain and with good radiation pattern at central frequency, but the bandwidth of both gain and impedance matching is obviously insufficient. That is resulted from the nature of a FPR resonator antenna. So, the next design steps should solve these problems.

C. Improved Design for Gain-Frequency Response



(a)



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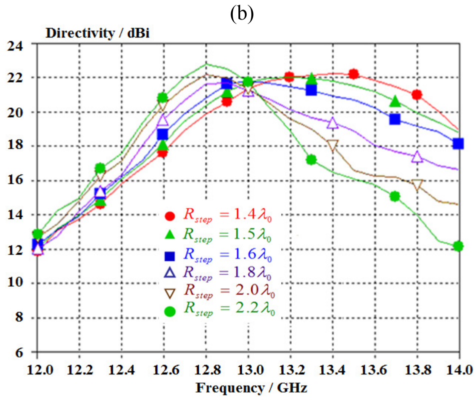


(c)

Fig. 4. Gain-frequency response design. (a) Diagram demonstration of resonant condition; (b) Vs. H_{step} (R_{step}=1.4 λ_0, H_2=0.49 λ_0); (c) Vs. $R_{step.}$ (H_{step}=0.04 λ_0, H_2=0.48 λ_0).

To enlarge the bandwidth of gain-drop, a key step is to flat the curve of gain-frequency response, based on reducing gradient difference between two sides of equality (1) denoted respectively as:

$$\Phi(f)=\phi_1(f)+\phi_2(f) \tag{3}$$

$$\Theta(f)=\frac{4\pi H}{c}f-2N\pi \tag{4}$$

Where reflection phase $\phi_1(f)$ from the grating-cover and $\phi_2(f)\equiv180^{\circ}$ from the metal-base had been given in Fig. 2; and N=0 usually for minimal thickness. So the resonant condition (1) can be simplified as

$$\phi_1(f)+\pi=4\pi Hf/c \tag{5}$$

also be demonstrated in diagram as Fig.4a.

The stepped base makes $H_1\neq H_2$ ($H_1<H_2$), corresponding to different resonant sub-frequencies (f_0^1 and f_0^2). Their difference $\Delta f=(f_0^1-f_0^2)$ can be adjusted by choosing $H_{step}=(H_2-H_1)$. Compared to the uniform base with only one f_0, Fig.4b shows that the gain-frequency response curve of stepped base becomes flatter obviously, and then the gain-drop bandwidth can be significantly enlarged when choosing a proper value of H_{step}. Once the stepped base with H_{step} is determined, the bandwidth of gain will also be changed with the radius of inner ground sub-area R_{step}, as shown in Fig.4c. A wide and flat gain-frequency response curve can be performed for a proper value of $R_{step}=1.4\lambda_0$. However, the performance of impedance matching is always dissatisfied without essential improvement, whatever the proper H_{step} and R_{step} are employed, which will be solved in the next design process.

D. *Improved Design for Impedance-matching*

The strong reflection returned into the feed seems due to direct reflection from the central part of cover, it may be dispersed by using an additional inverted metal conic structure with radius of R_{cone} and height of H_{cone} as shown in Fig. 5a, by referencing the experience in design a waveguide T-junction. Besides, the insertion length H_{wg} of waveguide plays as an additional adjusted parameter.

Fig. 5b,c,d display the influences of parameters $\{H_{wg}, R_{cone}, H_{cone}\}$ on the frequency response curves of antenna gain and S_{11}, respectively, under the case of other proper parameters from the above design procedure as: D_{grat}=10 mm, W_{grat}=4 mm, D_{grd}=140 mm, R_{wg}=8.3 mm, H_2=11.5 mm, H_{step}=0.7 mm, R_{step}=26.2 mm. Fig. 5b shows that the larger H_{wg} within $(0.05\sim0.15)\lambda_0$, the wider impedance bandwidth and the slightly lower peak gain.

Fig. 5c shows that R_{cone} affects the bandwidth of both impedance and gain obviously but contradictorily, which means a larger radius of cone is better for impedance bandwidth but results in narrower gain bandwidth and lower peak gain. However, Fig. 5d shows that impedance bandwidth can be improved significantly by tuning H_{cone}, while gain deteriorates very slightly. In a word, it is possible to achieve a wholly good performance for both impedance and gain bandwidth by an optimized design procedure.

III. PROTOTYPE WITH EXPERIMENTAL VERIFICATION

A prototype antenna was fabricated with optimized structural sizes: D_{grat}=10 mm, W_{grat}=4 mm, D_{grd}=140 mm (6 λ_0), R_{wg}=8.3 mm (0.36 λ_0), H_1=10.8 mm (0.47 λ_0), H_2=11.5 mm (0.5 λ_0), R_{step}=26.2 mm (1.1 λ_0), H_{wg}=3.6 mm (0.16 λ_0), R_{cone}=4.8 mm (0.21 λ_0), H_{cone}=6.0 mm (0.26 λ_0).

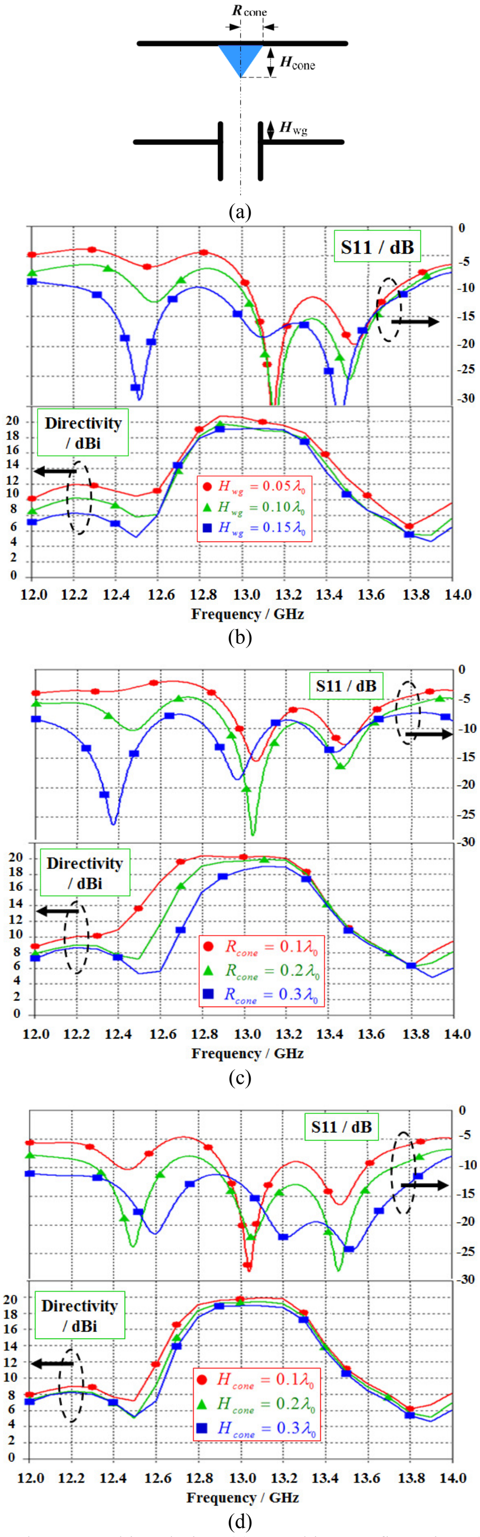
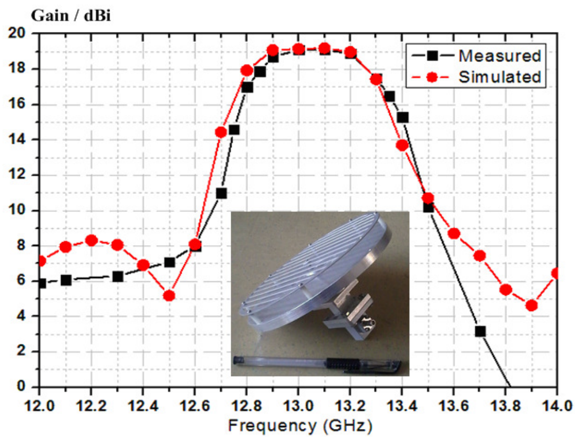
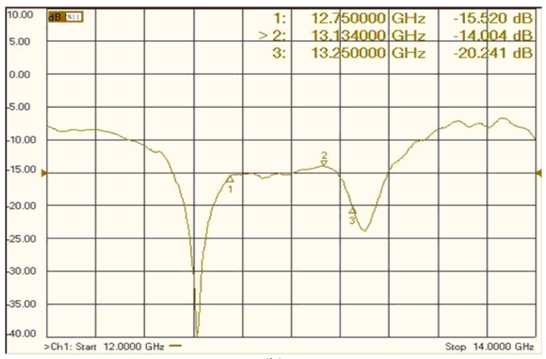


Fig 5. Impedance matching design. (a) Matching configuration; (b) *V*s. H_{wg} (R_{cone} =0.2 λ_0, H_{cone}=0.25 λ_0); (c) *V*s. R_{cone} (H_{wg}=0.15 λ_0, H_{cone}=0.25 λ_0); (d) *V*s. H_{cone} (R_{cone}=0.2 λ_0, H_{wg}=0.15 λ_0).

In Fig. 6a, the measured and simulated curves of gain-frequency response agree with each other very well; the peak gain approaches 19.2 dBi, bandwidth of 3-dB gain-drop covers (12.75~13.37) GHz; its band-pass filter-like curve of gain with distinct features of flat-top and abrupt-edges also benefits the system performance. Fig. 6b shows that measured return loss is better than −14 dB within the required frequency band.



(a)



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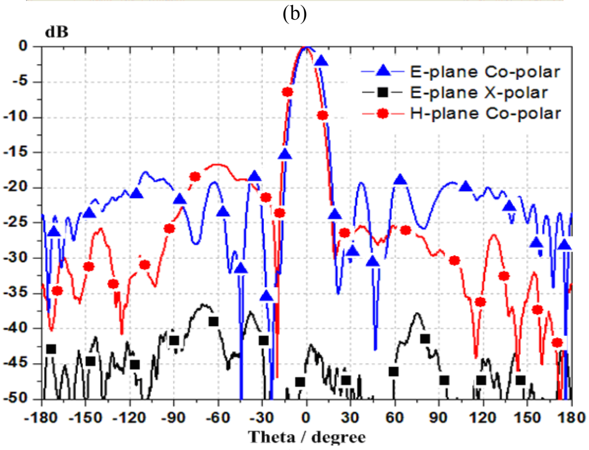

(c)

Fig.6. Prototype and performances. (a) Gain-frequency response; (b) Measured return loss; (c) Measured radiation patterns at 13.0 GHz.

Fig.6c presents the measured radiation patterns at designed frequency 13.0 GHz. The simulated ones are not shown for paper brevity. The simulated cross-polarization discrimination (XPD) is less than –40 dB within the main beamwidth in 45°-plane, and less than even –60 dB within whole angular range in both E- and H-plane. The measured XPD is also so good that the X-polar component in H-plane, which could not be shown in the figure's scale. The measured side-lobe level (SLL) is less than −15 dB in both E- and H-plane, which agrees well with the simulated ones. The radiation performances at the two edge frequencies have not been shown here for paper brevity, however they are quite similar except for a little higher SLL at high frequency.

IV. CONCULUSIONS

A novel structure of fully metallic CAFA antenna is proposed. Its design principle and simulation analysis are presented. A prototype for application of 13.0 GHz microwave radio-link has been designed, fabricated and measured for verification. Although the prototype antenna operates in a relative narrow frequency band, it can also be designed for the other wideband systems according to the above method. This kind of antenna exhibits low cost, high reliability, low profile, easy system integration and band-pass filter character. However, only about 20 dBi achievable gain and not consideration of radom are the main insufficiencies in this design, which would be improved in the future design by such as sub-array technique [3].

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