High Power Wideband Antennas for Electronic Warfare Systems

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HIGH POWER WIDEBAND ANTENNAS

FOR ELECTRONIC WARFARE SYSTEMS

by

Jaegeun Ha

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Department of Electric, Computer, and Energy Engineering

Dejan S. Filipovic

W. Neil Kefauver

The final copy of this thesis has been examined by the signatories, and we find that both the content and the form meet acceptable presentation standards of scholarly work in the above mentioned discipline.
This thesis researches several novel high power wideband antennas with omnidirectional, horizon- and broadside-directional, and bi-directional coverage for various electronic warfare (EW) systems. The developed antennas are designed and analyzed with the considerations in frequency domain (FD), time domain (TD), and power domain (PD) using full-wave electromagnetic simulations, transient analysis, multiphysics simulations, field analysis, equivalent circuit modeling, spherical mode decomposition, etc., as needed. The theoretical models of all antenna configuration are carefully validated with measurements.

Equivalent circuit models of monopole-like antennas are derived based on the physical structure and field distribution around a monopole. These models are used to ease physical understanding, analyze antenna impedance, and design derivatives with improved performance. A novel wideband omnidirectional antenna with consistent monopole-like patterns is proposed for millimeter-wave integrated towed decoy receiving subsystems by combining annular slot mode with a monopole into a single structure. The bandwidth of the combined antenna is improved to cover the wide range of radar signals from 18 GHz to 45 GHz, and its operating principle is analyzed using the proposed equivalent circuit model. In addition, compact paraboloid reflector fed by the combined antenna is designed for the same platform but with enabled horizon-directional coverage. The path toward using the proposed combined antenna for transient high power electromagnetics (HPEM) applications with good TD and PD performances is described.

Reconfigurability between steerable horizon-directional and omnidirectional modes for transient HPEM pulses is demonstrated by a monocone fed reflector antenna system. To do so, a conventional monocone is modified to a conical monopole first. The spherical mode expansion (SME) analysis shows that this structure has consistent monopole-like patterns and can be used as a feeder of a wideband reflector antenna system. Owing to the small diameter and consistent patterns of the modified monocone feeder, a paraboloid reflector is engineered and stable impedance and far-field performances from 1.66 GHz to 20
GHz are demonstrated. Analysis in TD and PD proves the suitability of the proposed antenna concept for transient HPEM applications. The electrical size of the monocone is also reduced by loading shorted semi-helical wires. The circuit model analysis demonstrates that typical miniaturization techniques require large monocone diameter though height can be greatly reduced. The proposed loading does not contribute to additional antenna volume, while lowering its electrical size by 34%, and the prototype monocone achieves good impedance match and gain from 1.23 GHz to 11 GHz.

Planar bi-directional and flush-mountable broadside unidirectional log-periodic (LP) antennas for long range communication, radar, or jammer systems with over a decade wideband performance are also introduced. To maintain inherent bi-directionality of a planar LP antenna, microstrip feed is preferred over a typical coax-based exciter. However, this leads to deteriorated impedance match and pattern distortion when the boom angle is not wide enough. To resolve this issue, a novel wide-boom geometry of the LP antenna is proposed. Since the wide-boom provides wide ground plane for the feeding microstrip line, good impedance match and gain performances can be achieved without sacrificing the antenna size. Moreover, electro-thermal multiphysics analysis of the antenna shows that the wide-boom also improves the temperature response, and thus, the average power handling capability is increased. The multiphysics simulation results are validated using a non-contact infrared (IR) camera. To achieve broadside unidirectional coverage, a slot-loaded cavity backing design is engineered. Instead of filling the cavity with carbon loaded absorber, the proposed slot-cavity backed LP antenna is loaded with the slot terminated by a coax cable that can be easily used for energy recycling. Fabricated prototype maintains high gain and efficiency with good impedance match over a wide bandwidth from 0.59 GHz to 5 GHz.

All ideas and concepts proposed in this thesis are thoroughly validated with circuit and full-wave simulations as well as experiments on representative prototypes.
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CHAPTER 1

Introduction

1.1 Motivation

Since the start of the new millennium, the electronic warfare (EW) technologies have received increased attention by defense ministries across the globe. These nonlethal technologies are needed to achieve control in the use of the scarce and critically important electromagnetic spectrum. EW systems can monitor RF spectrum, determine the direction or location and properties of various RF emissions, impede performance of radars or radar guided weapons, combat remote controlled improvised explosive devices, real-time assess the effect of jamming, and many other functions needed to achieve battlefield supremacy [1]. For example, some radar systems can be used to compare the transmitted signal with the reflected echo signal to determine the appearance of a hostile object or to guide a missile to the target [2]. To protect aircraft from such radar-guided missiles, a towed decoy system behind the aircraft first seduces and then misleads the missile’s on-board electronics to go after the decoy rather than the protected platform [3]. As an another example, the nonlethal directed energy weapon (DEW) technology uses the high power microwave radiation to temporarily disable or spoil electronic systems of unfriendly forces [4]. To enable unimpeded functionality of such EW systems, the performance of antennas including their cost, size, weight, and form factor is critically important.

It is often required that the EW antennas operate over a wide frequency range with good impedance match and consistent radiation patterns since radar systems can hop between different carrier frequencies [1]. Also DEW systems that use transient high power short pulses require antennas to have wide bandwidth in frequency domain (FD) and minimum distortion in time domain (TD) [4]. Furthermore, these antennas should be able to handle high peak and/or average power. In conjunction with these very general requirements, the antenna size needs to be as small as possible since they often dominate the size of EW systems and dictate the ways systems are being deployed. Small electrical size brings many fundamental issues in discussion (bandwidth, directivity, power handling, etc.), which makes the job of antenna
researchers and engineers even more daunting [5]. Therefore, the performance of EW antennas needs to be analyzed and assessed in frequency, time, and power domains, and significant improvements in their characteristics and capabilities are desired.

1.2 Background

As a part of electronic countermeasure (ECM) operation, towed radio decoy (TRD) systems have proven to effectively neutralize radar guided missiles, commonly operating through Ku-band [3]. Emergence of new technologies and systems has pushed the spectrum usage to higher frequencies so there is a need to extend the frequency range of TRDs. There are various ways how TRDs operate, but on common concept of operation is as follows: when a radar warning receiver on an aircraft detects radar signals from an enemy’s missile, the appropriate countermeasure waveform is created and sent to the decoy through a fiber cable. Within the decoy, this RF modulated light is converted back to RF, amplified, and backer-broadcast to missile by decoy’s transmitter antennas [6]. Consequently, radar cross section (RCS) of the decoy becomes larger than the aircraft so that the enemy’s missile is guided to the decoy instead of the aircraft. However, the complexity of such systems causes significant increase in cost to ensure they survive mechanical hardship and achieve desired RF performance. Moreover, the radar guided missiles are constantly improving and reach higher frequencies for which decoys may not have an appropriate answer. Thus, more capable but simplified and lower cost solutions that can reach ever increasing spectral needs are being sought.

As an effort to minimize the lethality during war operations while enhancing the effectiveness, nonlethal electronic attack (EA) systems manipulate high power electromagnetic (HPEM) pulse or directed energy using a high power capable antenna. HPEM systems are often classified into hypoband with bandwidth ratio \((br) < 1.01\), mesoband with \(1.01 < br \leq 3\), sub-hyperband with \(3 < br \leq 10\), and hyperband with \(br > 10\) [7]. While narrowband HPEM systems use a continuous wave (CW) or damped sinusoidal signals, (sub-) hyperband HPEM systems radiate short transient pulses with a high peak power. Therefore, sub-hyperband and hyperband antennas are required to have wide bandwidth in FD, minimum pulse
distortion in TD, and high peak power handling capability, whereas narrow band HPEM antennas need to have high average power handling capability.

Since the emergence of transient ultrawideband (UWB) technology in the 1980s, a myriad of UWB antennas were reported [8]. While commercial UWB communication systems are forced to comply with the low power emission limit [9], several UWB antennas are successfully adopted in wideband HPEM systems. One of the simplest forms is the transverse electromagnetic (TEM) horn that can achieve sub-hyperband to hyperband performance with good peak power handling capability and TD performance [10-12]. Also, modified forms of TEM horn antennas such as Shark, Vivaldi, and Scissors antennas were proposed in [13-15]. In addition, miniaturization of TEM horn antennas can be achieved by combining with the magnetic dipole mode [16-18].

A reflector antenna is also a suitable candidate for HPEM radiation. Since a reflector radiates collimated beams to a specific direction, a high gain and front-to-back ratio (FBR) can be achieved. While a narrowband feeder is typically used with a cylindrical parabolic or a dish-like paraboloid reflector, wideband flat or shaped reflector antennas fed by monopolar patches are also proposed for UWB communication and radar systems [19-22]. The most common antenna for a wideband HPEM radiation is the impulse radiating antenna (IRA). IRA is essentially a paraboloid reflector fed by a conical TEM transmission line [23-25]. Owing to the frequency independent (FI) properties of the feed and reflector structures, IRA can accomplish extremely wide bandwidth, minimum pulse distortion, and high power handling capability [11]. Detailed analysis and design procedure of IRA can be found in [26]. In [27], authors demonstrated IRA using a 57 cm paraboloid reflector that achieves 10 dB return loss bandwidth from 1.5 GHz to 13 GHz, therefore, its electrical size at the turn-on frequency is $2.85 \lambda_0 \times 2.85 \lambda_0 \times 1.14 \lambda_0$. Although this is not large compared to typical reflector antennas, much miniaturization and lighter weight are desired in practical circumstances considering the relatively low frequency range of transient HPEM applications [7]. IRA can be cut in half and placed on a ground plane to reduce the size, but it is still physically large [28]. Also, by having a proper resistive loading at the end of the conical transmission line, an IRA can achieve impedance match at very low frequencies where the electrical size of antenna is only
0.383 λ₀ × 0.383 λ₀ × 0.192 λ₀; however, gain is very low at low frequency band [29]. Moreover, for IRAs and even for TEM horns, horizontal scanning of the beam is prohibitive unless the entire HPEM system including a massive high voltage generator is rotated altogether. Such steering ability is desired for HPEM applications since target can be located or moved to any direction.

For narrowband HPEM or long-distance radar/communication systems with frequency hopping functionality, traditional FI antennas such as spiral, sinuous, and log-periodic (LP) antennas can be utilized owing to their extremely wide bandwidth in FD, light weight, and versatile form factors [30-32]. Moreover, FI antennas can accommodate multifunctional systems within their operating bandwidths. However, these antennas are often susceptible to the thermal breakdown. Such thermal failure can cause expansion and detachment of traces, change in the material property, melting dielectric, and structural deformation. In spite of possible hazards, the thermal performance of antennas and other RF devices has been seldom considered. In 1979, Bahl and Gupta analyzed the temperature rise in microstrip lines using the simplified parallel plate model [33]. Conduction and dielectric losses were calculated assuming no natural airflow and an isothermal ground plane. Although their analytical results offer a qualitative and quantitative understanding of the temperature rises, they are based on the oversimplified model and are limited to simple structures. Design of high power capable and complex antennas is often performed by following the rule of thumb guidelines such as use of wider and thicker strips, specialized substrates, and so forth. To assess the thermal performance of these antennas, laborious and expensive high power experiments are commonly used. Recent developments in multiphysics simulation tools have enabled numerical analysis of RF devices in multiple physical domains including electromagnetics, thermal, structural, and fluidal [34-37]. Based on this new capability, thermal and/or structural behaviors of RF devices were analyzed in [38-41]. However, thermal domain (ThD) performance of antennas has been barely considered.

1.3 Methodology

This thesis researches several antennas with added novel structural features for use in emerging wideband high power EW systems. These antennas enable coverage such as omnidirectional, horizon-
broadside-directional, and bidirectional. They are conceptualized, analyzed, designed, and then evaluated with considerations in frequency, time, and power domains. One of the goals desired is that the electrical size of each antenna system developed in this thesis be kept as small as possible and the instantaneous coverage with consistent near- and far-field performances be maintained over multi-octave wide bandwidths. Ansys HFSS, a commercial full-wave electromagnetic simulation software based on finite element method (FEM), is mainly used for frequency analysis [42]. Equivalent circuit model analysis is performed to gain insight in antenna impedance behavior, minimize mismatch loss, and reduce antennas’ electrical size. The circuit simulations are performed using NI AWR [43]. A spherical mode expansion (SME) code is utilized to analyze the radiated far-field of antennas [44, 45]. For TD analysis, antennas are modelled as multidimensional linear time invariant (LTI) systems and their transfer functions and impulse responses are used to calculate the output pulse from antennas [46-48]. When excited with high-power signals, antennas can experience breakdowns such as discharge through dielectrics, multipacting, and thermal heating. While dielectric and multipacting breakdowns are more critical for high peak transient pulses at low and high altitudes, respectively, the thermal breakdown likely happens before those for high average power CW and damped sinusoidal signals [49, 50]. The peak power handling capabilities of presented antennas are evaluated based on the electric field strength calculated in Ansys HFSS, and thermal analysis and measurement are performed using Ansys Workbench multiphysics simulation platform and a FLIR’s infrared (IR) camera [34, 51].

1. 4 Organization

This thesis is organized as follows:

- Chapter 2: wideband equivalent circuit model of monopole-like antennas with arbitrary dimensions is presented and empirical formulas for circuit parameters are determined. Since the proposed circuit model is derived based on the physical structure and field distributions of a monopole, it is successfully utilized to design an integrated matching element for a monopole and to analyze the combined antenna in Chapter 3. Also, the proposed circuit model is demonstrated to be valid for
annular slot and monocone antennas due to the equivalent radiation mode, and is used for reducing the electrical size of a monocone antenna in Chapter 4.

- **Chapter 3**: novel wideband monopole-like antenna sensors for millimeter-wave TRD receiving systems are introduced. Computational study detailing the impact of conducting cylinder on the performance of a monopole antenna is carried out, and simple, low-cost, and wideband connector antenna is used to exhibit the monopole antenna performance on a decoy-like cylinder. To further extend the monopole bandwidth, combination of a monopole mode with an annular slot mode in a single structure is discussed. For the sectorial coverage of TRD receiving systems, broadband offset paraboloid reflector antenna with the combined antenna feed is also presented. In addition, the combined antenna is evaluated in time and power domains to be considered for the possible use in transient HPEM applications.

- **Chapter 4**: a novel mechanically steerable reflector antenna with fixed monocone feed and miniaturization of a monocone are described. The monocone feed is designed using SME analysis and consistent monopole-like patterns over a wide bandwidth is achieved. Using the designed feeder, a wideband paraboloid reflector is engineered to miniaturize the electrical size while maintaining good performance in FD owing to the consistency and small diameter of the feeder. Also, the mechanical rotating system of the reflector with the fixed feeder is presented, facilitating the azimuthal scanning of the main beam for HPEM applications. Additionally, a novel miniaturization technique for a small diameter conical antenna is proposed and its reduced electrical size is followed with excellent wideband performance.

- **Chapter 5**: Bi- and broadside-directional (uni-directional) LP antennas operating over a wide instantaneous bandwidth are introduced. A novel wide-boom geometry is proposed to improve the performance of a planar bi-directional LP antenna without sacrificing its size. Theoretical basis and modeling of a microstrip patch antenna in electro-thermal domain using a multiphysics simulation tool is exploited to develop a more complex model and analyze the thermal behavior of the bi-directional LP antenna. Interesting correlation between the electrical and thermal behavior is seen.
and the modeling results are verified through a non-contact IR camera measurement. Finally, a novel slot-loaded cavity backing is introduced to transform the bi-directional patterns into uni-directional with a wide bandwidth and high efficiency demonstrated over almost a decade bandwidth.

- Chapter 6 provides concluding remarks and directions for the future work.
CHAPTER 2

Equivalent Circuit Model of Monopole-Like Antennas

2.1 Introduction

Modeling an antenna using circuit elements is an effective tool for design of matching circuits, co-simulation of front-end circuit components with antennas, calculating bandwidth, and so on [52, 53]. Since dipole/monopole antennas have been used from the very early days of radio, efforts from antenna engineers have produced many equivalent circuit models thereof. The simplest models are series and shunt RLC resonant circuits, but those are valid only near the series and parallel resonances, respectively. Chu presented a simple circuit model where each LC ladder network represents a specific spherical mode, but it is also valid in a narrow bandwidth [52, 54]. A broadband circuit model for a dipole antenna is first reported in 1981 using the dominant pole-pair approach based on Foster’s canonical form [55]. Also a modified form of [55] is proposed in [54] for electrically small antennas. In addition, typical impedance curve of a dipole is analyzed to derive a circuit model in a way that the pole at \( f = 0 \) is produced by a series capacitor, a series inductor in conjunction with the capacitor creates the series resonance, and full-wave resonances are generated by a shunt RLC circuit [56]. The bandwidth is further enhanced in [57] and empirical formulas for circuit parameters are given, so that one can build a circuit model of a dipole without simulations.

The majority of reported circuit models are derived from the typical impedance curve of a dipole. Despite good accuracy over a wide bandwidth, these circuit models are not well correlated with the physical structure of a dipole, while typical circuit models of antennas are derived based on physical insight [58-60]. Although most reported circuit models of a dipole begin with a series capacitor, the antenna structure does not contain such a series gap (capacitor). Because of this low correspondence of circuit model with the physical structure of the dipole, understanding impact of modifications in antenna structure on impedance is not straightforward.
This chapter presents a physically derived equivalent circuit model of a wire monopole with arbitrary dimensions. From the physical structure and field distribution, a monopole is modeled as a segment of transmission line terminated by a series RLC circuit as shown in Fig. 2.1. The agreement between full-wave and circuit simulations validates the model. Also, circuit parameters in the model are extracted for various antenna dimensions to obtain empirical formulas. In addition, the circuit model in Fig. 2.1 is used to model annular slot and monocone antennas since those have the equivalent fundamental mode of radiation. Empirical formulas for circuit parameters of those antennas are also given. The developed circuit model is utilized in subsequent chapters to analyze combination of two antenna elements, integrated matching element, and miniaturization of conical antennas.

This chapter is organized as follows:

- In Section 2.2, equivalent circuit model of a monopole and extracted circuit parameters are presented. Also, empirical equations are derived.
- In Section 2.3, circuit parameters and empirical equations for annular slot antenna are provided.
- In Section 2.4, circuit parameters and empirical equations for monocone antenna are provided.
- In Section 2.5, the chapter is summarized.

2.2 Monopole

Fig. 2.2 shows the analysis model of a coax-fed wire monopole on an infinite ground plane. The diameters of inner and outer conductors of the air-filled coax are \( D \) and 2.3\( D \), respectively, to maintain 50 \( \Omega \) characteristic impedance. The extension of center conductor by \( L \) functions as a wire monopole. The wire...
induces series inductance due to the circulating magnetic flux, and a shunt capacitance is formed between the monopole and the ground plane. The combination of these infinitesimal inductors and capacitors constitute a TEM transmission line segment as depicted in Fig. 2.3(a), where Poynting vector propagates along the monopole. Subsequently, a series inductor $L_A$ is placed in series as shown in Fig. 2.3(b), where strong magnetic field is formed. At the end of monopole, a strong electric field is formed as shown in Fig. 2.3(c) that indicates the capacitor $C_A$ is also connected in series. Then the radiation resistance $R_r$ completes the circuit model. Based on this physical observation, the equivalent circuit model for the monopole is derived as shown in Fig. 2.1.

Fig. 2.2. Analysis model of the coax-fed wire monopole antenna.

Fig. 2.3. Simulated (a) Poynting vector, (b) magnetic field, and (c) electric field of the monopole in elevation plane for $L = 0.25\lambda$. 

10
Fig. 2.4(a) and (b) show real and imaginary parts of the input impedance of the monopole, respectively, for various length-to-diameter ratio $L/D$. Full-wave simulation is performed first in Ansys HFSS and the equivalent circuit parameters are extracted using curve-fitting optimization tool in NI MWO. The circuit simulation data correlate very well with the full-wave simulation over two octave bands.

Fig. 2.4(a) and (b) show real and imaginary parts of the input impedance of the monopole, respectively, for various length-to-diameter ratio $L/D$. Full-wave simulation is performed first in Ansys HFSS and the equivalent circuit parameters are extracted using curve-fitting optimization tool in NI MWO. The circuit simulation data correlate very well with the full-wave simulation over two octave bands.

The extracted circuit parameters for various $L$ against log-scaled length-to-diameter ratio ($\log_{10}2L/D$) are shown in Fig. 2.5–2.8 and empirical equations are derived. Fig. 2.5 shows the extracted $Z_C$ and $R_r$ of the monopole. As $2L/D$ increases, the thinner the wire, the characteristic impedance and the radiation
resistance also increase, resulting in higher input resistance of antenna. Since $Z_C$ increases quadratically with $\log_{10} \frac{2L}{D}$ and is insensitive to $L$, the equation for $Z_C$ is obtained as

$$Z_C = 13.35 \cdot k^2 + 61.1 \cdot k + 2.45,$$

(2.1)

where $k$ is defined as

$$k = \log_{10} \left( \frac{2L}{D} \right)$$

(2.2)
for the sake of simplicity. The coefficients in the equation are obtained using least-square curve-fitting method. Also, $R_r$ increases logarithmically with $k$, thus

$$R_r = 92.1 - 90.32 \cdot 0.3^k. \quad (2.3)$$

Fig. 2.7. Extracted $L_A$ from the curve-fitting and empirical equations for various $L$ and $D$.

Fig. 2.8. Extracted $C_A$ from the curve-fitting and empirical equations for various $L$ and $D$.

The extracted $l_m$ is shown in Fig. 2.6. It is observed that $l_m$ quadratically decreases for $2L/D < 10$ and is almost constant for $2L/D > 10$. Thus, the empirical equation for $l_m$ is given as

$$l_m = L \cdot (0.41 + 4.33 \cdot 0.004^4). \quad (2.4)$$
Fig. 2.7 and 2.8 show the extracted $L_A$ and $C_A$. As expected, the thinner wire (larger $L/D$) provides the larger inductance and smaller capacitance. Since $L_A$ increases and $C_A$ decreases quadratically, the empirical formulas are given as

$$L_A = L \cdot \left( -20.2 \cdot k^2 + 196.5 \cdot k - 25.6 \right) \times 10^{-9},$$ \hspace{1cm} (2.5)$$

$$C_A = 26.6 \cdot L \cdot k^{-1.55} \times 10^{-12}. \hspace{1cm} (2.6)$$

A good agreement between formulas and circuit parameters is achieved, and the impact of the antenna dimensions on the circuit parameters is explicitly established.

![Analysis model of the open-ended coax annular slot antenna.](image1)

![Input impedance of the annular slot obtained from full-wave and circuit simulations.](image2)

**2.3 Annular Slot**

Since an annular slot antenna that has a constant magnetic loop current is equivalent to an electric monopole, the circuit model in Fig. 2.1 can be adopted to represent the impedance of an annular slot
antenna. Fig. 2.9 shows the analysis model of an annular slot antenna that is simply a coax open-ended at an infinite ground plane. Fig. 2.10 shows the input impedance of annular slot antenna against diameter $D$ normalized to the wavelength. As described in the previous section, the full-wave simulation is performed first, then the circuit parameters are extracted. A good agreement is achieved, thus, the same circuit model can be applied to represent the impedance of a monopole and an annular slot. As shown, the input resistance and reactance increase with frequency, and the antenna’s impedance behavior has high-pass nature.

![Graph showing input impedance against diameter](image)

Fig. 2.11. Extracted $Z_C$, $R_r$, and $l_m$ from the curve-fitting and empirical equations for various $D$.

Fig. 2.11 shows the extracted $Z_C$, $R_r$, and $l_m$ of the circuit model of annular slot antenna. As seen, $Z_C$ and $R_r$ are relatively constant regardless of $D$ for 50 $\Omega$ feed, and $l_m$ linearly increases with $D$. Thus, asymptotic equations for those parameters can be written as

$$Z_C = 61.2,$$  
(2.7)

$$R_r = 113.5,$$  
(2.8)

$$l_m = 0.32 \cdot D.$$  
(2.9)

Fig. 2.12 shows the extracted $L_A$ and $C_A$ that are proportional to $D$, thus asymptotic equations are simply,

$$L_A = 92.98 \cdot D \times 10^{-9},$$  
(2.10)

$$C_A = 8.73 \cdot D \times 10^{-12}.$$  
(2.11)

Owing to the simple structure of annular slot, the equations are relatively simple.
2.4 Monocone

The fundamental radiation mode of a monocone is identical to monopole and annular slot antennas; therefore, the circuit model is the same as in Fig. 2.1. Fig. 2.13 shows the analysis model of a monocone antenna on an infinite ground plane. Height $H$ and flare angle $\alpha$ are two independent variables of monocone. Fig. 2.14 shows the full-wave and circuit simulated impedance of monocone for various $\alpha$. Again, full-wave simulation is performed first and circuit parameters are optimized to be fitted, and good agreement is obtained.

![Diagram of monocone antenna](image)

Fig. 2.13. Analysis model of monocone antenna.
Fig. 2.15 shows the extracted $Z_C$ and $R_r$ in the circuit model of monocone. $Z_C$ decreases as $\alpha$ increases and the empirical formula is given as

$$Z_C = 54.76 \cdot \ln \left( \cot \left( \frac{\alpha}{4.299} \right) \right)$$

(2.12)

which is close to the well-known formula for the characteristic impedance of a conical transmission line,

$$Z_C \approx 60 \cdot \ln \left( \cot \left( \frac{\alpha}{4} \right) \right),$$

(2.13)
though coefficients are slightly modified. Also, $R_r$ exponentially decreases with $\alpha$, thus,

$$R_r = 301.9 \cdot \exp(-1.321 \cdot \alpha) + 25.13$$

(2.14)

The characteristic impedance of transmission line and the radiation resistance is insensitive to height of monocone, but are only functions of flare angle.

Fig. 2.15. Extracted $Z_C$ and $R_r$ from the curve-fitting and empirical equations for various $H$ and $\alpha$.

Fig. 2.16. Extracted $l_m$ from the curve-fitting and empirical equations for various $H$ and $\alpha$.

though coefficients are slightly modified. Also, $R_r$ exponentially decreases with $\alpha$, thus,

$$R_r = 301.9 \cdot \exp(-1.321 \cdot \alpha) + 25.13$$

(2.14)

The characteristic impedance of transmission line and the radiation resistance is insensitive to height of monocone, but are only functions of flare angle.

Fig. 2.16 shows the length of transmission line for various heights of the monocone. From the length of the edge of the monocone,
the empirical equation of $l_m$ can be obtained by modifying (2.15) as

$$l_m = 1.374 \cdot H \cdot \sec\left(\frac{\alpha}{2.412}\right) \quad (2.16)$$

Fig. 2.17 and 2.18 show the extracted $L_A$ and $C_A$, respectively. The larger the flare angle, the smaller the inductance and the larger the capacitance, just as thicker the monopole. The formulas for $L_A$ and $C_A$ are given in (2.17) and (2.18).
\begin{align*}
    L_a &= H \cdot \left(16.62 \cdot \alpha^2 - 111.9 \cdot \alpha + 172.1\right) \times 10^{-9}. \\
    C_a &= H \cdot \left(3.094 \cdot \exp(1.884 \cdot \alpha) + 9.693\right) \times 10^{-12}.
\end{align*}

2.5 Conclusion

In this chapter, equivalent circuit models for monopole-like antennas are presented. These models are derived from the physical structure and the specifics of the field distribution associated with a baseline monopole. The proposed circuit model for monopole is shown to be valid for annular slot and monocone antennas because the latter two have equivalent modes of radiation. It is demonstrated that the proposed model is valid over 4:1 bandwidth. Also, empirical formulas for circuit parameters are given so that one can calculate the circuit parameters of antennas with arbitrary dimensions. The impedance of dipole-like antennas can be obtained by simply doubling the impedance. The proposed circuit model will be used to design an integrated matching element for monopole antenna, to analyze the combined antenna, and to miniaturize monocone antenna throughout this thesis.
CHAPTER 3

Wideband Monopole-Like Antenna Sensors for Towed Airborne Decoys

3.1 Introduction

A current towed radio decoy (TRD) system with receiver and signal processor in the aircraft, transmitter and amplifier in the decoy, and fiber cable connecting the two can be simplified so that the receiver, transmitter, and processor are all integrated into a cylindrical platform as shown in Fig. 3.1(a). Herein, the radar signal is repeated, amplified, and delivered to the transmitter all within the decoy platform. An omnidirectional receiver antenna can be placed in the middle of the decoy platform. Although the system is greatly simplified, two transmitter antennas in fore and aft positions need to be turned on simultaneously in this scenario, which leads to the loss of more than a half of energy, since the direction of missiles may not be fully known. Thus, replacing an omnidirectional receiver antenna with two directional radiators as shown in Fig. 3.1(b) can improve the system efficiency because only one transmitter needs to be turned on upon which receiver detects the signal, though the system complexity is increased to some extent. In this chapter, three wideband omni-/horizon-directional receiver antennas are presented for such simplified TRD systems.

The TRD system is expected to be built in a conducting cylinder that considerably affects the antenna performance [61-63]. Therefore, understanding the effect of the cylindrical platform on the antenna performance is required. A fundamental study of monopole antenna performance on various cylindrical structures is presented. The surface currents on the cylinder induced by the antenna are computed and relevant physics is discussed to attain an insight on the impact of cylinder on antenna performance and the antenna isolation.

For omnidirectional coverage, monopole is the most widely used antenna owing to simple and low cost structure in a variety of applications such as cellular and ultra-wideband (UWB) communications, wireless body area networks (WBAN), radio frequency identification (RFID), just to mention a few [64-71]. At millimeter-wave frequencies, physically compact antenna can be readily designed; however, precise
and expensive fabrication process is often needed. Specific examples include on-chip inverted-F and quasi-Yagi antenna at 60 GHz manufactured with a back-end-of-line process and a surface micro-machined W-band CPW-fed vertical monopole antennas, respectively [72, 73]. Although these antennas have low profile and small size, their design is process-specific and manufacturing thereof is typically expensive. In this chapter, a very simple and low-cost implementation of a millimeter-wave wideband monopole antenna using an off-the-shelf connector is presented. The compact launch pin of the connector achieves 38% VSWR $\leq 2$ bandwidth with consistent radiation patterns. The connector antenna is fabricated, installed on a decoy-like cylinder, and its performance is demonstrated experimentally.

Since the frequency range of missile’s radar signals often falls in K/Ka-bands (18–40 GHz) and beyond (up to 45 GHz as reported), the receiver antenna of TRD also needs to cover such a wide bandwidth. While the bandwidth of realized connector antenna is from 24.7 GHz to 37.2 GHz, combination with annular slot mode is considered to further extend its bandwidth. Many monopole-based designs are reported to overcome its inherently narrow bandwidth. For example, printed monopole on a dielectric slab is shown to enhance the bandwidth [66, 67, 74-76]. However, inconsistent radiation patterns in the matched band and low polarization purity due to lateral currents often render this approach less practical. To address these

![Diagram](image.jpg)
issues, a dielectric resonator loading of a printed monopole is used [77-79]. At millimeter-waves, dielectrics may cause substantial gain reduction as in [80], so one needs to be very careful. In [81] and [82], a metallic plate is bent and rolled to achieve wideband impedance match with vertically polarized omnidirectional radiation. A sub-wavelength metamaterial coating for a wire monopole is suggested in [83] to produce additional resonance that extends the bandwidth. An alternative to the previous works that is arguably simpler, lower loss/cost, and easier to scale to lower/higher frequencies is proposed herein. The approach integrates thick vertical monopole with annular slot in a single structure named the combined annular slot-monopole antenna (CASMA).

The annular slot, also known as the circular diffraction antenna, can be implemented as an open-ended coaxial transmission line [68, 84]. However, this implementation is not widely used due to the small radiation resistance within the single-mode frequency band of the coaxial feed line. Since a vertically oriented electric monopole has radiation pattern similar to a horizontally oriented magnetic loop (annular slot), the integration of the two concepts in a single antenna structure may extend the impedance bandwidth while maintaining consistent radiation patterns. This concept is implemented herein to achieve wideband impedance match with VSWR ≤ 2 from 17.9 GHz (high-pass) and stable radiation pattern up to 48 GHz. To analyze this hybrid/combined structure qualitatively, impedance and radiation properties of monopole and annular slot antennas are investigated first with the help of equivalent circuit model derived in the previous chapter. Then the combined antenna is analyzed and its equivalent circuit model is also derived.

For the sectorial coverage illustrated in Fig. 3.1(b), the omnidirectional antenna needs to be replaced by a directional aperture that keeps the back radiation very low (i.e. high front-to-back ratio (FBR)). This chapter also presents a broadband compact paraboloid reflector design with CASMA feed, in which the size and shape are designed with electrical and application-driven considerations. Though a reflector for a monopole antenna usually operates efficiently only in a narrow bandwidth due to the destructive interference [85], the stable performance of the reflector antenna over the whole decoy bandwidth is achieved by tuning the position of focal point of reflector.
The performance of CASMA in time domain (TD) and power domain (PD) is also evaluated to be considered for transient high power electromagnetics (HPEM) applications. To do so, CASMA is scaled down to 1–3 GHz range and impedance match is improved by adding a ring matching element. In addition, the transition from commercial 50 Ω coax to the antenna aperture is modified to enhance the peak power handling capability. As a result, the designed antenna has achieved good performance in frequency, time, and power domains over a 3:1 bandwidth.

This chapter is organized as follows:

- Section 3.2 studies the performance of a monopole on a cylindrical platform.
- Section 3.3 introduces a simple wideband connector antenna and its performance on a decoy-like cylinder.
- Section 3.4 examines impedance and far-field properties of monopole and annular slot antennas.
- Section 3.5 proposes the CASMA for omnidirectional TRD receiver applications.
- Section 3.6 discusses the paraboloid reflector with CASMA feed for directional TRD receivers.
- Section 3.7 considers the performance of CASMA in power and time domains for transient HPEM systems.
- Section 3.8 concludes the chapter.

### 3.2 Antennas on Cylindrical Platform

Antenna performance, especially the radiation characteristics, is largely dependent on its platform. In this section, effects of a cylindrical platform on antenna performance are studied. The simulation setup is depicted in Fig. 3.2. A 2.5 mm long vertical wire monopole is positioned on a conducting cylinder (resonant at 30 GHz). The diameter and the length of the cylinder are parameterized as $D$ and $L$, respectively, and the effects of those parameters are evaluated.

The radiation patterns of the wire monopole antenna for various cylinder dimensions are shown in Fig. 3.3. As seen, the typical monopole-like patterns in E-planes and nearly omni-directional patterns in H-plane are observed. It is noticed that beamwidth in the xz-plane is much wider than that in the yz-plane and
beamwidth in the yz-plane decreases as $L$ increases. Also, the thicker cylinder causes the narrower beamwidth and the maximum radiation closer to the horizon in the xz-plane. Increased length and decreased diameter lead to higher cross-pol gain in the xy-plane. Overall, these patterns indicate that the monopole maintains its typical far-field performance when integrated on a cylindrical platform, but high cross-pol gain is induced.

![Diagram of antenna setup](image)

Fig. 3.2. Illustration of the simulation setup for a vertical wire antenna on a cylindrical platform.

![Radiation patterns](image)

Fig. 3.3. Radiation patterns of the monopole in (a) xz-plane, (b) yz-plane, and (c) xy-plane on the cylinder with various dimensions. Black: co-pol, gray: cross-pol.

### 3.2.1 Parametric Study

Larger datasets depicting the antenna performance as the cylindrical platform varies are illustrated in Fig. 3.4-3.7. The dimensions of the cylinder are normalized to the free space wavelength (1 cm at 30 GHz).
Also the diameter is log-scaled for the xz-plane and the length is log-scaled for the yz-plane. Fig. 3.4(a) and (b) show the half-power beamwidth (HPBW) in xz- and yz-planes, respectively. For small diameter cylinders, the HPBW is around 100°, while it is 60–80° when the diameter is larger than one wavelength. The length merely affects the xz-plane pattern. On the other hand, the length is the dominant factor in yz-plane. As the length increases from 1 to 45 wavelengths, the HPBW in the yz-plane changes from 60° to 10°. This narrow beamwidth is consequence of the main beam radiating at the edges of the cylinder due to the surface currents when the cylinder is electrically long. Note that the impedance of antenna is not affected much by cylindrical dimensions.

Fig. 3.4. Contour plots of HPBW for various dimensions of the cylinder in (a) xz-plane and (b) yz-plane.

Fig. 3.5. Contour plots of the beam squint (tilt elevation angle) of the main beam with respect to the horizon for various cylinder dimensions in (a) xz-plane and (b) yz-plane.
The tilt of the main beam due to finite size ground plane is referred to as tilt elevation angle. Fig. 3.5 shows this angle with respect to $z = 0$ plane. In the $xz$-plane, the beam squint is small and does not vary a lot. In the $yz$-plane, the beam squint is about $30^\circ$ for electrically short cylinders, and $10^\circ$ for long ones. This effect of cylinder’s dimensions is observed in the peak gain as shown in Fig. 3.6. In the $xz$-plane, since the diameter is electrically as small as $0.5–5$ wavelengths, the peak gain changes from $0.5$ dBi to $3$ dBi while the nominal peak gain of a monopole on an infinite ground plane is about $6$ dBi. In the $yz$-plane, the $6$ dBi peak gain is obtained when the length is $10–20$ wavelengths. In the longer ($>20 \lambda_0$) cylinder regime where the edge radiation dominates, the peak gain can be as high as $12$ dBi. Edge radiation is also stronger as the diameter is reduced. Thus, the surface currents on long and thin cylinders noticeably increase the antenna gain.

Fig. 3.6. Contour plots of the peak gain for various cylinder dimensions in (a) $xz$-plane and (b) $yz$-plane.

Fig. 3.7. Contour plots of the (a) WoW and (b) the cross-pol gain for various cylinder dimensions in $xy$-plane.
Fig. 3.7(a) shows the wobble of the wave (WoW), a measure of omnidirectionality in the H-plane (xy-plane). A very thin cylinder \((D < 0.5\lambda_0)\) has 4 dB WoW, while other cylinders have WoW below 3 dB. The cross-pol gain in the H-plane is shown in Fig. 3.7(b). The thinner and the longer cylinder give rise to the higher cross-pol. In this study, vertical-pol is the co-pol and the horizontal-pol is the cross-pol. From this observation and the radiation pattern in Fig. 3.3, one can find that the edge radiation from the cylinder is comparable to the y-oriented dipole antenna that contributes to the high co-pol gain in the yz-plane and high cross-pol gain in the xy-plane.

3.2.2 Coupling

The cylinder dimensions are crucially important not only due to the effective size of the ground plane, but also due to the possibility of exciting a horizontal dipole moment with the surface currents supported by the cylinder. Furthermore, this current may cause increased coupling between multiple antennas on the cylinder when more than two antennas are implemented to achieve dual-polarized receiving. To evaluate the cross-talk (coupling/isolation) between multiple monopole antennas on a cylinder, we created a simulation setup as shown in Fig. 3.8. The cylinder length and diameter are 18” and 1.875”, respectively, to replicate a virtual TRD system. In Fig. 3.8(a), three wire monopoles are located along the circumference, and in Fig. 3.8(b), four wire monopoles are along the length.

![Fig. 3.8. Simulation model of multiple wire monopole antennas on the cylinder. (a) Monopoles along the circumference. (b) Monopoles along the length.](image)

The electric field distribution on the cylinder at 30 GHz induced by the Ant#1 is shown in Fig. 3.9. In the contour plot, the field strength is illustrated on a log-scale. As seen, the field strength rapidly decays along the circumference due to diffraction on the curved surface. However, the induced field can propagate
through the length of the cylinder with very low resistive losses. Thus, when the surface wave reaches the edge, the field strength is abruptly reduced, implying that the strong radiation at the edge occurs. Note that this behavior can also be observed at the other frequencies.

Fig. 3.9. Magnitude of the electric field distribution on the cylinder at 30 GHz produced by a wire monopole antenna placed in the middle of the cylinder.

Fig. 3.10. Magnitude of the electric field (a) along the circumference and (b) along the length.

The magnitudes of electric field along the circumference and the length of the cylinder at different frequencies are shown in Fig. 3.10(a) and (b), respectively. The induced field exponentially decreases along the circumference and the strengths are quite similar for different frequencies. On the other hand, the field strength along the length becomes smaller as frequency increases due to increased radiation from surface wave. Fig. 3.11(a) shows the S-parameters of the monopole antennas along the circumference (see Fig. 3.8(a)). The return losses of all three antennas are higher than 10 dB from 23.4–35.0 GHz. The isolation
between the 90° offset antennas is 40–50 dB and it exceeds 55 dB with the antennas on the opposite side of the cylinder (180°). At 30 GHz the difference between the S21 and S31 is 23 dB since the difference between the strength in the induced field is 111.6 V/m. Fig. 3.11(b) shows the S-parameters of the monopoles along the length (see Fig. 3.8(b)). The isolations are 35–40dB, 40–50dB, and 50–60dB for monopoles at (0, 3”, 0.9375”), (0, 6”, 0.9375”), and (0, 9”, 0), respectively.

![S-parameters of monopole antennas on the cylinder (a) along the circumference and (b) along the length.](image)

3. 3 Connector Antenna

A simple implementation of millimeter-wave monopole antenna is presented in this section. The presented antenna achieves wide impedance bandwidth and consistent radiation patterns with low cost/loss structure. Also, the antenna is evaluated on a decoy-like cylindrical platform.

3. 3. 1 Bandwidth of a Monopole

Fig. 3.12 shows the 10 dB return loss bandwidth and resonant frequency of a 2.5 mm length wire monopole on an infinite ground plane for various $l/d$ values. The gap feeding is used in this study. Since the length of the wire is fixed at 2.5 mm, the resonant frequency is maintained around 28–29 GHz. As expected, the 10 dB bandwidth is increased as $l/d$ decreases. The widest bandwidth achieved for this monopole is 38% when $l/d = 8$, and for the smaller $l/d$, bandwidth is decreased due to the increased capacitance between the bottom of wire and the ground plane [86].
3.3.2 Realization and Measurement

To realize the electrically thick wire monopole antenna, a commercially available 2.4 mm connector is used as shown in Fig. 3.13(a). The connector is the core of a 0.5” x 0.5” antenna platform. A launch pin with diameter \( d = 20 \) mil (0.0508 \( \lambda_0 \) at 30 GHz) is inserted into the socket and trimmed to have length \( l/2 = 2.3 \) mm (0.23 \( \lambda_0 \) at 30 GHz). The length to diameter ratio \( (l/d) \) is \(~9.1\); therefore, the pin functions as an electrically thick vertical monopole when configured as seen in Fig. 3.13(a). In Fig. 3.13(b), the connector socket is recessed into the antenna platform and only the launch pin is above the ground plane.

Fig. 3.12. 10 dB return loss bandwidth and resonant frequency of a wire monopole antenna on an infinite ground plane as \( l/d \) varies from 5 to 1000.

Fig. 3.13. Photographs of 2.4 mm connector and its launch pin used as wire monopole antenna. (a) The standalone connector antenna. (b) Antenna installed on a finite ground plane.
Fig. 3.14 shows the VSWR of the connector antenna on the circular ground plane. The measured 2:1 VSWR bandwidth covers 24.7–37.2 GHz. As seen the simulations and the measurements have excellent correlation. Fig. 3.15(a) and (b) show the measured co- and cross-polarized radiation patterns in the E- and H-planes, respectively. In addition, the realized gain and the WoW are shown in Fig. 3.16. As seen, the gain is higher than 4.5 dBi and stable throughout the matched bandwidth. Moreover, good H-plane pattern omnidirectionality and cross-polarization rejection of over 25 dB are measured. The WoW is below 3 dB. Higher WoW in measurement is resulted from the imperfect cutting of the ground plane and the positioning
of the antenna at the center. These measurements clearly show that the proposed connector antenna has consistent radiation pattern in both E- and H-planes over its full bandwidth.

![Graph showing realized gain and WoW vs frequency](image)

**Fig. 3.16.** Realized gain and WoW.

![Connector antenna installed on a decoy-like conducting cylinder](image)

**Fig. 3.17.** The connector antenna installed on a decoy-like conducting cylinder.

### 3.3.3 Connector Antenna on Cylinder

The connector monopole antenna is installed on a decoy-like cylinder as shown in Fig 3.17. The connector socket is recessed into the cylinder, so only the 2.3 mm length launch pin is above the cylinder. Fig. 3.18 shows the VSWR of the antenna on the cylinder. Since the cylinder does not alter much of the impedance, the VSWR is almost the same as on the circular ground plane. The radiation patterns in xz-, yz-, and xy-planes are shown in Fig 3.19(a), (b), and (c), respectively. A consistent radiation across the bandwidth is observed. The xz-plane pattern is comparable to a monopole antenna on a small ground plane. In yz-plane, the edge radiation due to the surface current along the cylinder causes the narrow main lobe at
$\theta = 82^\circ$. This edge radiation results in bumps along $\pm y$-axis in the $xy$-plane radiation pattern and degrades the omnidirectionality as shown in Fig. 3.20.

Fig. 3.18. VSWR of the connector antenna on the decoy-like cylinder.

Fig. 3.19. Measured radiation patterns in (a) $xz$-, (b) $yz$-, and (c) $xy$-planes. Solid: co-pol, dashed: cross-pol.

Fig. 3.20. Elevation angle in $xz$- and $yz$-planes and WoW in $xy$-plane.
3.4 Single Mode Antenna Elements

Impedance and far-field properties of wire monopole and annular slot antennas are studied in this section and will be used to design a wideband CASMA in the subsequent section.

3.4.1 Monopole

A baseline model of a monopole antenna used for analysis is illustrated in Fig. 3.21. Electrically thick wire on an infinite ground plane is fed by a 50 Ω coax. The length \( L \) is fixed at 3 mm while the diameter \( D \) is swept from 1 mm to 2.5 mm. Because the high capacitance between the wire and the ground plane deteriorates the impedance match for thick wires [86], the gap \( G \) is increased with increased \( D \) (\( G = D/5 \)).

![Fig. 3.21. A thick wire monopole antenna on an infinite ground plane.](image1)

![Fig. 3.22. Equivalent circuit model of the monopole.](image2)

<table>
<thead>
<tr>
<th>D (mm)</th>
<th>1.0</th>
<th>1.5</th>
<th>2.0</th>
<th>2.5</th>
</tr>
</thead>
<tbody>
<tr>
<td>( Z_C ) (Ω)</td>
<td>56.2</td>
<td>41.6</td>
<td>34.4</td>
<td>31.2</td>
</tr>
<tr>
<td>( l_m ) (mm)</td>
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<td>1.68</td>
<td>2.11</td>
<td>2.79</td>
</tr>
<tr>
<td>( L_A ) (pH)</td>
<td>360</td>
<td>288</td>
<td>174</td>
<td>154</td>
</tr>
<tr>
<td>( C_A ) (fF)</td>
<td>111</td>
<td>155</td>
<td>269</td>
<td>295</td>
</tr>
<tr>
<td>( R_r ) (Ω)</td>
<td>57.4</td>
<td>47.2</td>
<td>39.3</td>
<td>34.4</td>
</tr>
</tbody>
</table>

Table 3.1. Equivalent circuit parameter values for monopole.
Fig. 3.22 depicts the equivalent circuit model of a monopole from Fig. 2.1 and Table 3.1 lists the circuit parameters for various $D$. As $D$ increases, $Z_C$, $L_A$, and $R_r$ decrease and $l_m$ and $C_A$ increase as expected from Chapter 2. Fig. 3.23 shows the impedance of the monopole for various diameters simulated in HFSS and computed from the circuit model for extracted parameters listed in Table 3.1. As seen, good agreement is achieved over 15–50 GHz band. As expected, the impedance variation is reduced as the diameter is increased, so the bandwidth is enhanced. However, the matched (to 50 $\Omega$) bandwidth is limited when the

Fig. 3.23. Impedance of a thick monopole for various diameters obtained from full-wave and circuit simulations.

Fig. 3.24. Simulated VSWR of the monopole for various diameters.
diameter is ‘too large’ because of the small nominal resistance resulted from the decreased $Z_C$ and $R_r$. Fig. 3.24 shows the full-wave simulated VSWR of the monopole. As seen, the VSWR ≤ 2:1 bandwidth for $D = 1.5$ mm is slightly wider than that for $D = 1.0$ mm; however, the bandwidth is decreased when $D$ is further increased.

Fig. 3.25. Annular slot as an open-ended coax. (a) Configuration. (b) Magnetic current distribution.

<table>
<thead>
<tr>
<th>$D$ (mm)</th>
<th>1.0</th>
<th>1.5</th>
<th>2.0</th>
<th>2.5</th>
</tr>
</thead>
<tbody>
<tr>
<td>$Z_C$ (Ω)</td>
<td>72.1</td>
<td>60.4</td>
<td>65.2</td>
<td>63.0</td>
</tr>
<tr>
<td>$l_m$ (mm)</td>
<td>0.35</td>
<td>0.44</td>
<td>0.62</td>
<td>0.77</td>
</tr>
<tr>
<td>$L_A$ (pH)</td>
<td>100</td>
<td>164</td>
<td>154</td>
<td>221</td>
</tr>
<tr>
<td>$C_A$ (fF)</td>
<td>7.83</td>
<td>12.8</td>
<td>17.8</td>
<td>22.0</td>
</tr>
<tr>
<td>$R_r$ (Ω)</td>
<td>110</td>
<td>103</td>
<td>111</td>
<td>113</td>
</tr>
</tbody>
</table>

Table 3.2. Equivalent circuit parameter values for annular slot.

### 3.4.2 Annular Slot

The baseline model of an annular slot antenna is illustrated in Fig. 3.25(a). This is simply an open-ended 50 Ω coaxial transmission line [87]. The inner and outer diameters of the coax are $D$ and $2.3D$, respectively, with the air filling between. The opening of the coax is shaped into an annular slot, supporting a constant magnetic loop current as shown in Fig. 3.25(b). The magnetic loop current is calculated from the simulated electric field using HFSS field calculator based on Huygen’s equivalence principle (3.1) [88, 89].

$$\overline{M_s} = -2\hat{z} \times \overline{E}. \quad (3.1)$$

As seen, the magnetic loop current is constant along the circumference and inversely proportional to the radius. This magnetic loop antenna is equivalent to an electric monopole; therefore, the equivalent circuit model is identical to that of monopole shown in Fig. 3.22. The obtained circuit parameters are shown in Table II. Fig. 3.26 shows the simulated impedance of the annular slot antenna for various diameters. As with monopole, the circuit simulation data agree well with those obtained by the full-wave simulation. The
Antenna resistance is increased for larger diameters, whereas the reactance is closer to zero for larger slots mainly due to the increased $l_m$. Simulated VSWR of the annular slot for various diameters is shown in Fig. 3.27. The impedance match of annular slot has high-pass response, thus, an extremely wide impedance bandwidth can be achieved. Although this flush-mountable annular slot antenna can achieve a very broad impedance bandwidth, the matched bandwidth is well above the higher order mode cutoff frequency of the corresponding coax, preventing its widespread use in practical applications.

Fig. 3.26. Impedance of the annular slot for various diameters obtained from full-wave and circuit simulations.

Fig. 3.27. Simulated VSWR of the annular slot for various diameters.
The radiation pattern of this constant magnetic loop antenna is theoretically predicted as

$$\left| E_\theta \right|^2 = \left[ A_0 \cdot J_1 \left( ka \cdot \sin \theta \right) \right]^2$$

(3.2)

where $E_\theta$ is the vertical component of electric field intensity, $A_0$ is the normalization constant, $k$ is the wavenumber, $a$ is the effective radius of the loop, and $J_1$ is the Bessel function of the first kind. Effective radius $a$ can be approximated to the average of inner and outer radii of the coax.

$$a = \left( \frac{D}{2} + \frac{2.3D}{2} \right) / 2 \approx 0.825D$$

(3.3)

Fig. 3.28 shows the calculation of (3.2) (dashed lines) and the simulated radiation pattern of the annular slot antenna in Fig. 3.25(a) (solid lines) over 30–100 GHz range. The diameter $D$ in the simulation is 3 mm.
and effective radius $a$ in calculation is 2.47 mm. The correlation between the two confirms (3.2) and (3.3).

Also, the radiation patterns resemble those produced by electric monopole at low frequencies, but null and side lobes at horizon begin to rise as the effective radius exceeds half-wavelength ($\gtrsim$60GHz) [84, 90].

3.5 Combined Annular Slot-Monopole Antenna (CASMA)

The impedance variation of a monopole can be reduced by increasing its diameter (i.e. decreasing the length-to-diameter ratio); however, the matched bandwidth is reduced due to the decreased nominal resistance. On the other hand, the input resistance of an annular slot antenna is increased with the diameter, but highly capacitive reactance impedes the impedance match under the fundamental mode of operation. By exploiting above discussed impedance properties, a wideband impedance match can be achieved with CASMA. Moreover, since the radiation characteristics of monopole and annular slot are equivalent to each other, a consistent omnidirectional pattern can be maintained over a large part of impedance bandwidth.

3.5.1 Antenna Design

The analysis model showing the integration of electric monopole and magnetic loop modes into CASMA is illustrated in Fig. 3.29. The baseline is grounded 50 $\Omega$ coaxial line with its central conductor extended above the ground plane. In addition to the typical electric monopole mode, the coax-fed wire monopole configuration supports magnetic loop antenna mode in the annular slot. The radiation pattern produced by this magnetic current resembles that of electric monopole when radius of the loop is below half-wavelength.

Fig. 3.29. Model of a combined annular slot-monopole antenna (CASMA).
Fig. 3.30. Equivalent circuit model of CASMA.

<table>
<thead>
<tr>
<th>$D$ (mm)</th>
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<th>1.5</th>
<th>2.0</th>
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</tr>
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<tbody>
<tr>
<td>$Z_{C1}$ ($\Omega$)</td>
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<td>50.9</td>
<td>53.1</td>
<td>56.7</td>
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<tr>
<td>$l_{m1}$ (mm)</td>
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<td>0.29</td>
<td>0.43</td>
<td>0.45</td>
</tr>
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<td>$L_{A1}$ (pH)</td>
<td>237</td>
<td>543</td>
<td>611</td>
<td>510</td>
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<tr>
<td>$C_{A1}$ (fF)</td>
<td>42.2</td>
<td>51.8</td>
<td>48.8</td>
<td>60.6</td>
</tr>
<tr>
<td>$R_{r1}$ ($\Omega$)</td>
<td>228</td>
<td>146</td>
<td>116</td>
<td>115</td>
</tr>
<tr>
<td>$Z_{C2}$ ($\Omega$)</td>
<td>107</td>
<td>105</td>
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<td>66.4</td>
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</tbody>
</table>

Table 3.3. Equivalent circuit parameter values for CASMA.

Fig. 3.31. Simulated impedance of CASMA for various $D$.

Fig. 3.30 shows the equivalent circuit model of CASMA to give an insight into how two modes are combined. Due to the mutual orientation between monopole and annular slot, two series RLC circuits with transmission lines, representing annular slot and monopole, are connected in parallel. Table 3.3 summarizes
CASMA’s circuit parameters for various $D$ obtained by the same procedure described above. Note that the values are modified from single mode operation parameters given in Tables 3.1 and 3.2 due to the interaction between the two modes.

The simulated impedance of CASMA is shown in Fig. 3.31, and good agreement between the full-wave and circuit simulations is obtained as before. As the diameter increases, the impedance variation is reduced as expected; however, the input resistance is greater than that of a monopole and the reactance is more capacitive owing to the contribution from the annular slot mode. Fig. 3.32 shows the VSWR of the CASMA for various $D$. As seen, the impedance bandwidth (VSWR $\leq 2:1$) is enhanced for larger diameter due to the fact that the bandwidth of annular slot mode approaches that of monopole when combined, whereas two modes are decoupled when the diameter is small (e.g. $D = 1.0$ mm). In fact, CASMA’s bandwidth features high-pass nature much like annular slot.

![Simulated VSWR of the CASMA for various D.](image)

Fig. 3.32. Simulated VSWR of the CASMA for various $D$.

Fig. 3.33 shows the simulated horizon gain of CASMA on an infinite ground plane. Since the radiation pattern of annular slot mode is ‘squeezed’ when the loop radius is larger than half-wavelength [90], the horizon gain begins to decrease at higher frequencies. To avoid this undesirable pattern inconsistency and keep VSWR below 2:1 over 18–45 GHz, $D = 2$ mm is chosen.
Effects of the annular slot mode on the bandwidth of monopole can be demonstrated by changing the slot diameter while keeping the monopole dimensions fixed. In Fig. 3.34, the monopole diameter is 2 mm, and a ring-shaped cap is placed on the annular slot to ‘partially’ close the opening. In this analysis, the slot diameter is reduced to $D_{AS}$ and varies from 3.4 mm to 4.6 mm. The impedance of the annular slot mode changes, but the characteristic impedance of the feeder and the input impedance of the wire are almost unchanged. As shown, the antenna bandwidth is reduced as $D_{AS}$ decreases, while the bandwidth of the

Fig. 3.33. Simulated horizon gains of CASMA for various $D$.

Fig. 3.34. Simulated VSWR of the coax-fed wire monopole with different diameter of annular slot $D_{AS}$.

Effects of the annular slot mode on the bandwidth of monopole can be demonstrated by changing the slot diameter while keeping the monopole dimensions fixed. In Fig. 3.34, the monopole diameter is 2 mm, and a ring-shaped cap is placed on the annular slot to ‘partially’ close the opening. In this analysis, the slot diameter is reduced to $D_{AS}$ and varies from 3.4 mm to 4.6 mm. The impedance of the annular slot mode changes, but the characteristic impedance of the feeder and the input impedance of the wire are almost unchanged. As shown, the antenna bandwidth is reduced as $D_{AS}$ decreases, while the bandwidth of the
monopole mode is kept at 18–35 GHz. This is not surprising since the smaller $D_{AS}$ leads to decoupling between annular slot and monopole modes. Conversely, the larger $D_{AS}$ brings the bandwidth of annular slot mode closer to the monopole bandwidth; therefore, the impedance bandwidth is improved.

Fig. 3.35. CASMA configuration and its constitutive parts.

3.5.2 Fabrication and Measurement

Fig. 3.35(a) shows the fabricated CASMA configuration. Careful implementation strategies are adopted to realize simple and low-cost millimeter-wave antenna structure. The monopole with conically tapered transition is fabricated using CNC machining. Its lower and upper diameters are 0.508 mm and 2 mm, respectively. The monopole is then placed in a commercial off-the-shelf 2.4 mm connector that is rated up to 50 GHz. The assembly is completed with a cylindrical block having a conical hole. This conical coax transition increases the diameter of the annular slot and thus controls the operating bandwidth of the annular slot mode, while minimizing the possible effects of higher order modes. The diameter of the conical hole is 2.3 times larger than that of the conical wire, so the 50 Ω impedance is maintained throughout this
transition. Fabricated CASMA is assembled on a 203 mm diameter (12.2 \( \lambda_0 \) at 18 GHz) circular ground plane as shown in Fig. 3.35(b). The size of ground plane is larger than 10 \( \lambda_0 \) over the entire bandwidth so that its effect on radiation pattern is insignificant.

![Fig. 3.36. VSWR of the CASMA for various heights of the transition \( H \).](image)

Fig. 3.36. VSWR of the CASMA for various heights of the transition \( H \).

![Fig. 3.37. VSWR and realized peak gain at approximately 18° elevation of CASMA on 203 mm diameter ground plane.](image)

Fig. 3.37. VSWR and realized peak gain at approximately 18° elevation of CASMA on 203 mm diameter ground plane.

Fig. 3.36 shows simulated VSWR of the CASMA for different heights of conical transition \( H \). As seen, once the height is sufficient (~3 mm in this case), its further increase has minimal effect on electrical performance. To ease the fabrication and keep low cost, the height \( H \) is set as 5 mm.
Simulated and measured VSWR and realized gain are shown in Fig. 3.37. The measured VSWR is below 2:1 from 17.9 GHz to 50 GHz while the simulated values are slightly higher at high frequencies. For comparison, the simulated VSWR of a monopole from Fig. 3.21 with $D = 2$ mm is also included. As seen, its bandwidth is noticeably narrower than CASMA, therefore, confirming bandwidth improvement due to the annular slot mode. The measured realized gain remains stable throughout the measured frequency range with nominal value of 5.5 dBi.

![Fig. 3.37: Measured normalized radiation patterns of the CASMA in (a) xz- and (b) xy-planes. Solid: co-pol, dashed: cross-pol.](image)

Fig. 3.38. Measured normalized radiation patterns of the CASMA in (a) xz- and (b) xy-planes. Solid: co-pol, dashed: cross-pol.

![Fig. 3.39: Peak-gain elevation (lift-off) angle in E-plane and WoW in H-plane.](image)

Fig. 3.39. Peak-gain elevation (lift-off) angle in E-plane and WoW in H-plane.
The measured radiation patterns of the fabricated prototype at four frequencies within 20 GHz to 50 GHz in xz- and xy-planes (E- and H-planes) are shown in Fig. 3.38(a) and (b), respectively. The consistent radiation patterns over a broad bandwidth are obtained indicating that the two modes coherently contribute to the antenna far-field. The cross-polarization rejection of more than 20 dB is also achieved. It is noticed that the direction of the maximum gain in E-plane is elevated at 50 GHz as the higher order mode is developed. The direction of peak gain in E-plane elevated from the horizon is depicted in Fig. 3.39. Due to the finite size of the ground plane, the main beams are elevated by 10° to 18°. From 48 GHz, the elevation angle is increased to above 40° as expected. The WoW in the H-plane is also shown in Fig. 3.39. As seen, the measured WoW is below 2.5 dB over the entire bandwidth, further demonstrating that the two modes are coherently added in far-field.

3. 6 CASMA-Fed Wideband Offset Paraboloid Reflector

For directional coverage of a TRD receiver shown in Fig. 3.1(b), directional radiation patterns with high FBR are required over the decoy bandwidth (18–45 GHz). CASMA is a good candidate as a feeder of a wideband reflector antenna owing to the stable impedance and radiation characteristics. Using CASMA as feeder, a wideband paraboloid reflector is designed.

3. 6. 1 Phase Center Offset of CASMA

For design of the reflector, the phase center location of the feeder and variation thereof are of crucial importance. Since a monopole on a ground plane has the same radiation characteristic with that of a dipole in free space, the phase center of monopole is expected to be at the feed. However, for a wideband monopole, such as electrically thick connector antenna or CASMA, the phase center varies against frequency owing to the higher order modes. For example, when the length of a monopole is three quarters of a wavelength, the current maximum occurs at two thirds of the monopole above the feed and the phase center should coincide with the current maximum. This phase center variation is numerically calculated and the reflector is designed based on this property.
Fig. 3.40. Simulation setup in HFSS to find phase center of CASMA. The center of hemispherical radiation boundary moves up along z-axis from the origin (feeding point), and the phase of the radiated electric field is calculated at the boundaries. When the center of hemisphere coincides with the phase center of the antenna, the phase maps of electric fields at each boundary would be constant. Keeping this in mind, we calculated the peak to peak phase variation of the electric field within $\theta = 0–45^\circ$ for various offset values at different frequencies. At each frequency, the value of offset that minimizes the phase variation is the phase center of the antenna. Fig. 3.41 shows the numerically calculated
phase center of the antenna against frequency. As seen, the phase center moves up along the z-axis from 0 mm to 2 mm as frequency increases.

To take the phase center variation into account for the reflector design, the focal point of a paraboloidal reflector is moved and the reflector is cut at the ground plane. Fig. 3.42(a) and (b) show VSWR and peak gain of CASMA fed paraboloid reflector antenna with the focal point on the ground plane (half-cut reflector), and Fig. 3.42(c) and (d) show those with focal point 2 mm above the ground plane (offset paraboloid). Focal distance $F_0$ is swept from 3.0 mm to 4.0 mm. As expected, turn-on frequency and in-band mismatch loss of both antennas are improved for larger $F_0$ owing to the less interaction between the feeder and reflector. Also, when $F_0$ is small, impedance match of half-cut reflector above 30 GHz is deteriorated, although electrical distance between feeder and reflector is farther at high frequencies. This is because radiation from CASMA with moved phase center at high frequencies is reflected back to the feeding transmission line. The impedance match can be improved by simply increasing $F_0$; however, a
significant gain drop occurs at high frequencies due to the destructive interference. By moving the focal point of reflector by 2 mm as shown in Fig. 3.43(c) and (d), both impedance match for small $F_0$ and gain drop for large $F_0$ at high frequencies are much improved since the phase center of feeder approaches the focal point of reflector. At low frequencies, although phase center of feeder is underneath the focal point of reflector, the reflected wave does not penetrate into the feeding line, so the good impedance match is maintained.

Fig. 3.43. Parametric study for focal height and distance of paraboloid reflector represented by contour plots of (a) maximum VSWR, (b) minimum FBR, (c) gain flatness, and (d) minimum gain within 18–45 GHz. (e) Null gain at 45 GHz. (f) Overlapped plot. Design goals are denoted as a contour curves with arrows.
3. 6. 2 Parametric Study

The antenna performance can be evaluated by various criteria and proper trade-off is often sought. To find the design parameters that critically impact the overall antenna performance, we perform an extensive parametric study. Fig. 3.43(a-d) show the contour plots of maximum VSWR, minimum FBR, gain flatness, and minimum gain of the CASMA within 18–45 GHz for various focal heights in z-axis and focal distances in y-axis of the paraboloid. The subtended angles in both E- and H-planes are kept 90°. From Fig. 3.43(a-d), minimizing VSWR, maximizing FBR, and flattening the gain are achieved simultaneously by increasing focal distance. However, this also gives rise to a null in the main lobe. Fig. 3.44 shows the radiation pattern in E-plane at 45 GHz for various $F_0$ while the focal height is fixed at 2 mm. As seen, the undesirable null appears around $\theta = 70^\circ$ due to the destructive interference and becomes deeper as $F_0$ increases. Let us define the gain at $\theta = 70^\circ$ at 45 GHz as ‘null gain’ (higher null gain is preferred). Fig. 3.43(e) shows the null gain for various focal points. Increase in focal distance decreases the null gain; therefore, a compromise for the parameters of paraboloid has to be found based on this tradeoff relationship. The desired focal distance and height of the reflector can be found by overlapping the criteria curves in a single plot as shown in Fig. 3.43(f). Our design goals include: VSWR below 2.5:1, FBR higher than 20 dB, minimum gain above 8 dBi, gain flatness below 4.5 dB, and null gain higher than 5 dBi. In Fig. 3.43(f), the desired dashed region is found as we choose 3.5 mm focal distance and 2.4 mm height offset.

![Fig. 3.44. Radiation pattern in E-plane at 45 GHz for various focal distances.](image)

Moreover, the subtended angle and the height of the paraboloid can be determined in the same manner. Fig. 3.45 shows the contour plots of parametric study on subtended angle and height of paraboloid where the focal distance and offset height are fixed as discussed. Lower VSWR, higher FBR, higher minimum
gain, and higher null gain are obtained as height increases. Also a much flatter gain is attained for smaller subtended angle. The design goals depicted in Fig. 3.45(f) are: VSWR < 2.5, FBR > 20 dB, minimum gain > 8 dBi, gain flatness < 3.5 dB, and null gain > 6 dBi. Notice that more rigorous criteria can be satisfied. The dashed area denotes the set of design parameters that result in good overall performance. Based on this, we choose 8 mm as the height and 80° as the subtended angle.

Fig. 3.45. Parametric study for subtended angle and height of paraboloid reflector represented by contour plots of (a) maximum VSWR, (b) minimum FBR, (c) gain flatness, and (d) minimum gain within 18–45 GHz. (e) Null gain at 45 GHz. (f) Overlapped plot. Design goals are denoted as contour curves with arrows.
3.6.3 Measurement

The fabricated reflector with the CASMA feed is measured on a 190 mm diameter circular ground plane as shown in Fig. 3.46. The reflector is manufactured using CNC machining and placed on the ground plane using a screw. Fig. 3.47 shows VSWR and the realized gain of the antenna. The measured VSWR is below 2.2:1 and gain is higher than 10 dBi from 18 GHz to 45 GHz. The good impedance match is maintained despite the close proximity between the feeder and the reflector because of the small diameter of the feeder. In addition, the stable gain curve is obtained owing to the nearly constant pattern from the feeder.

Fig. 3.46. Fabricated offset paraboloid reflector.

![Fabricated offset paraboloid reflector](image)

Fig. 3.47. VSWR and realized gain of the designed reflector antenna.

![VSWR and realized gain graph](image)
FBR and elevation angle are shown in Fig. 3.48. The designed reflector antenna measures FBR > 20 dB though the electrical size of the reflector is only $0.60\lambda_0 \times 0.48\lambda_0 \times 0.20\lambda_0$. Also, a stable elevation angle over the decoy bandwidth is attained. Fig. 3.49(a), and (b) show the measured radiation patterns in E-, and H-planes, respectively. Stable horizon-directional radiation patterns are observed in both planes due to the consistency of the feeder.

Fig. 3.48. FBR and elevation angle of the designed reflector antenna. Discretization of beam peak is due to data density.

Fig. 3.49. Measured radiation patterns in (a) E- and (b) H-planes.
3.7 CASMA for Transient HPEM Systems

The proposed CASMA and the reflector thereof can be easily scaled to lower and higher frequency bands to be used for the other wideband applications. For example, CASMA with/without the reflector can be possibly used as a transient HPEM radiator with omnidirectional and horizon-directional pattern reconfigurability. To examine such a possibility, CASMA is scaled to lower frequency band of 1–3 GHz, and time and power domain characteristics are demonstrated in this section.

Fig. 3.50 shows the scaled version of CASMA in 1–3 GHz range. The height of monopole is 54 mm and the diameter of annular slot is 69 mm. As shown in the previous sections, increase in monopole diameter $M_D$ results in enhanced impedance bandwidth, but decreased gain bandwidth. Herein, $M_D = 30$ mm is chosen. As expected, CASMA can cover a 3:1 bandwidth in FD.
3. 7. 1 Impedance Match in Power Domain

Impedance match is one of the key antenna performance parameters in FD. Most communication systems require antennas to have VSWR better than 2 over the frequency band of interest, while receiving antennas often have relaxed requirement of VSWR better than 2.5 or 3. For high power transmitting antennas, however, more rigorous criterion in impedance match is often required since the reflected wave degrades overall system efficiency and can damage the pulse generator. To illustrate this point, the reflection coefficient $\Gamma$ is calculated in terms of VSWR.

$$|\Gamma| = \frac{VSWR - 1}{VSWR + 1}$$  \hspace{1cm} (3.4)

Using (3.4), VSWR of 2 gives $|\Gamma| = 0.33$, meaning that one third of the input pulse voltage can be reflected back to the signal generator when the maximum in the power spectral density (PSD) of input pulse coincides with the frequency of maximum VSWR. This reflection can be suppressed by further improving the impedance match. For instance, VSWR of 1.25 yields $|\Gamma| = 0.11$ so that the reflected pulse is less than one ninth of the input.

In addition, the reflected wave limits the power handling capability of the antenna feed network due to the excited standing wave. The maximum and minimum voltages of the standing wave along the feed line are calculated as

$$V'_{\text{max}} = |V'| (1 + |\Gamma|),$$  \hspace{1cm} (3.5)

$$V'_{\text{min}} = |V'| (1 - |\Gamma|).$$ \hspace{1cm} (3.6)

Equation (3.5) indicates that the peak value of electric field can be increased by a factor of $(1+|\Gamma|)$, therefore, the peak power handling capability is limited by

$$\text{PHLF} = 20 \cdot \log_{10} (1 + |\Gamma|) \text{ [dB]},$$  \hspace{1cm} (3.7)

where PHLF stands for power handling limitation factor. The reflection coefficient ($|\Gamma|$ and $S_{11} = |\Gamma|^2$) and PHLF for various VSWR are listed in Table 3.4. In order to avoid power handling limitation, the impedance match of an antenna has to be as good as possible.
3.7.2 Ring Matching Element

As seen in the previous subsection, CASMA has VSWR below 2.2 over 1.0–3.5 GHz range. Owing to the high-pass nature of the annular slot mode, the impedance match is maintained through 10 GHz and beyond; however, impedance match above 3 GHz is meaningless since the pattern becomes inconsistent even if the connector maintains a single mode excitation. In this subsection, we outline a simple approach to further reduce in-band mismatch loss of CASMA, which leads to better total efficiency and power handling capability.

In Fig. 3.31, it is shown that the reactance of CASMA is capacitive for the most of bandwidth that is attributed to the contribution from the annular slot mode. To improve the impedance match, this capacitance needs to be compensated. This can be accomplished by adding an inductive element. Fig. 3.51 shows the equivalent circuit model of the proposed concept where CASMA is inductively coupled to an inductor and a parasitic capacitor. This concept is realized by a ring atop of the monopole as shown in Fig. 3.52. The ring is coupled to the monopole through the gap between, and the height of the ring induces an inductance. In addition, a parasitic capacitance is formed between the ring and the ground plane. Fig. 3.53(a) and (b).
depict the simulated magnetic field distributions with and without the ring, respectively. As shown, the magnetic field is higher around the ring, indicating increased inductance.

Fig. 3.52. Configuration of matching ring around the wire monopole.

Fig. 3.53. H-field around CASMA (a) with and (b) without the ring.

Fig. 3.54. (a) Impedance and (b) VSWR of the CASMA with matching ring for various gaps $S_G$. 

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Parametric studies to determine the dimensions of the matching ring are performed. Fig. 3.54(a) and (b) show the impedance and VSWR of the antenna, respectively, for various gaps between the ring and the CASMA, $S_G$. Those without the ring are also included for comparison. As seen, CASMA without the ring has capacitive reactance especially at 2.5 GHz. This capacitance is compensated by the inductively coupled ring and the reactance increases as $S_G$ is increased. As a result, VSWR is much improved in 1–3 GHz range. Also, it is seen that the ring causes an unwanted self-resonance formed by $C_R$ and $L_R$ at 3–3.5 GHz range, which reduces the impedance bandwidth though in-band match is improved. To avoid the self-resonance approaching to the operating bandwidth of the antenna (1–3 GHz), $S_G = 1$ mm is chosen.

Fig. 3.55. (a) Impedance and (b) VSWR of the CASMA with matching ring for various heights $S_H$.

Fig. 3.56. (a) Impedance and (b) VSWR of the CASMA with matching ring for various widths $S_W$. 

Parametric studies to determine the dimensions of the matching ring are performed. Fig. 3.54(a) and (b) show the impedance and VSWR of the antenna, respectively, for various gaps between the ring and the CASMA, $S_G$. Those without the ring are also included for comparison. As seen, CASMA without the ring has capacitive reactance especially at 2.5 GHz. This capacitance is compensated by the inductively coupled ring and the reactance increases as $S_G$ is increased. As a result, VSWR is much improved in 1–3 GHz range. Also, it is seen that the ring causes an unwanted self-resonance formed by $C_R$ and $L_R$ at 3–3.5 GHz range, which reduces the impedance bandwidth though in-band match is improved. To avoid the self-resonance approaching to the operating bandwidth of the antenna (1–3 GHz), $S_G = 1$ mm is chosen.
The effects of the height of the ring $S_H$ on impedance and VSWR are shown in Fig. 3.55(a) and (b), respectively. Again, the self-resonant frequency of the ring is decreased as the height increases while in-band match is improved. To obtain the flat impedance curve and low VSWR over a wide bandwidth, $S_H = 35$ mm is chosen. The impedance and VSWR variation for changing the width of the ring $S_W$ are shown in Fig. 3.56(a) and (b), respectively. We choose $S_W = 7$ mm to maximize VSWR $\leq 1.25$ bandwidth. Note that the ring does not affect the far-field performance except at the self-resonant frequency.

![Diagram](image)

Fig. 3.57. Side view of the proposed CASMA with matching element, N-connector feed, and the conical transition between them.

### 3.7.3 High Power Transition Design

A coaxial transmission line, in which the inner and outer diameters are the same with those of the monopole and annular slot, respectively, are used in the aforementioned computational studies for the sake of simplicity. However, such a large coax not only contains higher order modes in the operating bandwidth of the antenna, but also it is not compatible with commercial connectors. Therefore, a transition from a commercial connector to the antenna is designed. An N-connector feed is used herein owing to the high power handling capability and standardized dimensions. Note that the transition presented in this subsection can be readily modified for any other connectors. Fig. 3.57 shows the configuration of the proposed...
CASMA with the ring, N-connector feed, and the conical transition between them. The inner and outer dimensions of the used N-connector launch are 3.0 mm and 10.3 mm, respectively, and the medium is filled with PTFE ($\varepsilon_r = 2.2$) [91]. The transition is essentially an air-filled ($\varepsilon_r = 1$) conical coaxial transmission line. Since the permittivities of filling medium in the transition and the connector are different, either inner or outer diameter has to be discontinuous to attain continuous characteristic impedance of 50 $\Omega$. The continuous outer diameter is preferred, otherwise the wave propagation from the N-connector is somewhat blocked by the outer conductor of the transition. Thus, the inner and outer diameters of transition at the bottom are chosen to be 4.8 mm and 10.3 mm, respectively, and those at the top are 30 mm and 69 mm, respectively. This makes smooth transition from PTFE-filled N-connector to air-based CASMA design, while minimizing the effect of higher order modes. The height of the transition $T_H$ is determined based on a parametric study as shown in Fig. 3.58. As seen, $T_H = 60$ mm provides widest VSWR $\leq 1.25$ bandwidth.

Fig. 3.58. VSWR of the antenna for different transition lengths $T_H$.

Power handling capability of the HPEM CASMA antenna is likely limited by the corona breakdown at low altitude rather than thermal breakdowns since the antenna is mostly metallic. At high altitude, multipacting breakdown is more critical [49], but the antenna is assumed to be used at low altitude herein. Thus to evaluate the power handling capability of the antenna, the peak electric field intensity is computed. Fig. 3.59 shows the maximum electric field intensity $|E|$ of the antenna. The antenna is divided into three
regions: connector, conical transition, and monopole. As seen, the highest electric field is formed in the connector region because of high permittivity of PTFE. However, since the PTFE has higher dielectric strength of 20–173 MV/m than air (3 MV/m) [92], the breakdown in connector can be disregarded. Then the breakdown will likely happen at the transition region filled with air. Fig. 3.60 shows the electric field distribution at the transition from PTFE-filled N-connector to air-filled conical coax. It can be seen that the peak electric field occurs at the beginning of the inner conductor of the conical coax because of the discontinuity.

![Graph of electric field intensity vs. frequency](image1)

**Fig. 3.59.** Maximum electric field intensity of the antenna within connector, transition, and monopole regions for 1 W input.

![Electric field distribution](image2)

**Fig. 3.60.** Electric field distribution in the transition and the connector region.
The strong electric field sparks due to the discontinuity can be eliminated by making both inner and outer conductors continuous. In the meantime, the characteristic impedance should be kept 50 Ω throughout the transition. This can be accomplished by partially filling the conical coax region with PTFE as illustrated in Fig. 3.61. The inner and outer diameters at the bottom of the conical coax are the same with those of N-connectors, and the PTFE fulfills the bottom surface. Also, to keep the monopole dimensions the same, the portion of PTFE continuously decreases to zero at the top. As a result, no abrupt change in inner and outer conductors is made, but also the characteristic impedance is maintained at 50 Ω throughout the transition.

Fig. 3.61. The proposed transition conical coax, partially filled with PTFE.

Fig. 3.62. Maximum electric field intensity within connector, conical coax, and antenna regions with and without the partial PTFE filling.
Fig. 3.62 shows the maximum electric field intensity of the antenna in three regions for partially PTFE-filled transition and all-air-filled transition. The peak electric field at the antenna region is not changed much, but that in the connector and transition regions are decreased, which enhances the power handling capability of the antenna. Though the ambient air condition is assumed in the transition, much higher power handling is attainable with pressurized gas or high vacuum [50].

Fig. 3.63. Configuration of the proposed antenna with the optimized ring fed by N-connector through the conical transition.

### 3. 7. 4 Time Domain Performance

Having designed the conical transition and feed network, the matching ring is slightly modified to have the optimum impedance match. The dimensions of the ring element are optimized using generic algorithm optimization tool in HFSS as shown in Fig. 3.63. The optimization goal is to minimize the VSWR within 1–3 GHz. The HFSS simulated VSWR of the proposed CASMA with ring matching, fed by an N-connector through the conical transition is shown in Fig. 3.64. The VSWR without the matching ring is also shown for comparison. As seen, the VSWR without the ring is below 2.36:1 from 0.94 GHz to 3.5 GHz. With the ring, the VSWR is below 2:1 from 0.91 GHz to 3.45 GHz, and below 1.25:1 from 1.20 GHz to 2.99 GHz.

To evaluate the TD performance of the antenna, Fig. 3.65 shows the simulated input pulse, reflected pulse, and radiated pulse at the broadside (100 mm away from the antenna) with and without the ring. The
simulation is performed using HFSS transient solver. The input pulse is the second order Gaussian pulse with 10 dB spectrum bandwidth from 0.68 GHz to 3.34 GHz. The shapes of the radiated pulses with and without the ring are both good replications of the input pulse, indicating a good TD performance of the antenna. Additionally, the reflected pulse without the ring varies from -0.19 V to +0.19 V, while that with the ring are within -0.12–0.06V; thus reduced reflection is demonstrated.

Fig. 3.64. VSWR of the proposed antenna with and without the ring.

Fig. 3.65. Input, reflected, and radiated pulsed of the antenna with and without the ring.
3.8 Conclusion

In this chapter, wideband omni- and horizon-directional receiving antennas for TRD systems are presented. Since the decoy platform is cylindrical, a computational study is performed to understand the impact of cylindrical dimensions on radiation patterns, which are characterized in three principal planes using normalized contour plots. In the xz-plane, the increase in diameter results in narrower beamwidth, smaller elevation angle, and higher peak gain. In the yz-plane, the increase in length of the cylinder also gives rise to narrower beamwidth, smaller elevation angle, and higher peak gain. A high gain above 6 dBi is caused by the edge radiation from the surface current on the cylinder. This surface current is stronger for thinner and longer cylinders. In the xy-plane, the edge radiation brings about high cross-pol gain. Furthermore, the magnitude of surface current on the cylinder can be used to predict the antenna isolations on the cylinder when multiple antennas are installed on a single cylindrical platform. The surface current distribution on the decoy-like cylinder and its impact on antenna isolation are also presented.

The antenna performance on a decoy-like cylinder is demonstrated by a wideband coaxial connector antenna. Using the electrically thick connector pin, the wideband monopole can be easily tuned to a quarter wavelength at the desired frequency within the operational bandwidth of a chosen connector. The wideband impedance match with the consistent radiation patterns, moderate gain, and stable elevation angle are demonstrated. Moreover, the antenna bandwidth is maintained on the cylindrical platform and stable radiation performance is observed.

To further enhance the bandwidth of a monopole, the magnetic loop current of annular slot mode is combined with the monopole mode. The equivalent circuit models for monopole, annular slot, and the combined antennas are developed to provide better physical understanding. The fabricated CASMA operates from 17.9 GHz to 48 GHz with VSWR ≤ 2, nominal realized gain of 5.5 dBi, and consistent monopole-like patterns in both E- and H-planes. The inherent practical features of CASMA and excellent electrical performance make this antenna viable candidate for omnidirectional TRD receiver applications.

To attain horizon-directional pattern, a wideband paraboloid reflector design is developed. Based on the phase center variations of the monopole antenna feed, the focal center of a paraboloid reflector is offset.
As a result, lower VSWR and higher gain at upper band are achieved. In addition, the destructive interference between the direct and reflected waves is mitigated. The design parameters of the reflectors are found by using contour plots of specific performance parameters. The CASMA-fed offset paraboloid reflector achieves VSWR $\leq 2.2:1$, FBR above 20 dB, gain $> 10$ dBi, and consistent radiations patterns over 18–45 GHz while the electrical size is only $0.60\lambda_0 \times 0.48\lambda_0 \times 0.20\lambda_0$. Therefore, the designed antennas are suitable for TRD receiver subsystems with omnidirectional or directional coverage.

CASMA is also considered for transient HPEM radiators. The impedance match of the low frequency version of CASMA is improved by adding a ring matching element in order to alleviate the power handling limitation due to the standing wave. In addition, the high power conical coax transition from standard N-connector to the antenna is designed. The designed antenna achieves VSWR $\leq 1.25:1$ from 1.20 GHz to 2.99 GHz, gain higher than 1.6 dBi, good omnidirectionality, minimized reflected pulse, and consistent radiation pattern.
CHAPTER 4

Small Wideband Omnidirectional and Directional Antennas

for Transient HPEM Applications

4.1 Introduction

A mechanically steerable wideband paraboloid reflector antenna with a fixed omnidirectional feed is introduced. The developed configuration with fixed feed allows reflector to freely rotate, thus the horizontal scanning of the beam becomes rather easy. In addition, by using an omnidirectional feed, the antenna performance is maintained the same during the beam scanning. Finally, the radiation can be reconfigured into a steerable in azimuth directional beam with unfurled reflector and omnidirectional mode with recessed reflector.

The designed feeder belongs to the monocone family, thus allowing wide bandwidth, stable phase center, good time domain (TD), frequency domain (FD), and power domain (PD) characteristics, and circular symmetry. To ensure stable patterns, a 90° monocone is modified to a conical monopole where in the upper region of the cone is trimmed to be cylindrical. The 2 dBi bandwidth framework suggested in [93], where the minimum gain in the field of view (FOV) from 51° to 90°, is higher than 2 dBi, is used to define the monopole-like pattern bandwidth. The height of the conical junction is determined to maximize the 2 dBi bandwidth and the design process is simplified using the spherical mode expansion (SME) technique.

Using the designed feeder, a wideband paraboloid reflector is engineered to miniaturize the electrical size while maintaining good performance in FD and TD. Then, a mechanically steerable reflector antenna subsystem is designed and fabricated. The measurements show good impedance match and gain, high FBR, and directional horizon beams over a wide bandwidth. The proposed antenna is also characterized in TD and PD. Excellent agreement with theory indicates the robustness of the proposed concept.
The miniaturization of omnidirectional conical antenna is also discussed in this chapter. Broadband omnidirectional antennas with compact size are needed for commercial, medical, and military applications. In 1976, Goubou reported electrically small monopole antenna with integrated matching elements that to date remains a gold standard in terms of size/bandwidth [94]. The bandwidth of a monopole can also be increased by using top hat loading and shorted pins which produce additional parallel resonance [95, 96]. This approach has seen many different implementations recently including, for example, a sectorial loop antenna composed of a conical patch and shorting arches [97]. Although compact size and wide bandwidth are achieved, its circular asymmetry degrades omnidirectionality. Shorted top hat loaded body of revolution (BOR) antennas with better omnidirectionality can be found in [98-101]. Common to the above mentioned works is focus on reducing the antenna height rather than diameter. While electrical height at the lowest operating frequency can be shorter than 0.1 λ₀, the diameter is typically larger than 0.3 λ₀, making those antennas not suitable for the feeder of an electrically small reflector. Furthermore, the hyper-bandwidth of a primitive conical antenna is often shrunk to around 3:1. In [102], to maintain pattern consistency and return loss bandwidth, two half-diamond loops are arranged to radiate at different frequency ranges. Since the smaller radiator is embedded into the larger, the antenna height is kept almost the same. Despite the low-profile and consistent patterns, a complex feed network is required and the mismatch loss is relatively high.

A novel semi-helical loading is proposed here to miniaturize the electrical size of a modified monocone without increasing its diameter. The parallel inductance is controlled by the helical turn ratio; thus the turn-on frequency is reduced by ~30% for the fixed antenna size. The developed equivalent circuit model is used to better understand the miniaturization effects. The theory is experimentally verified and good impedance match, consistent gain around 5 dBi, and omnidirectional radiation over a wide bandwidth are seen.

This chapter is organized as follows:

- Section 4.2 describes a modified monocone feeder to achieve consistent monopole-like patterns over a wide bandwidth
- Section 4.3 discusses design of an electrically small mechanically steerable reflector antenna.
Section 4.4 proposes a semi-helical loading for the miniaturization of a monocone.

Section 4.5 concludes the chapter.

4.2 Modified Monocone with Consistent Patterns

To begin with, let us examine the radiation properties of monopole and monocone antennas over a wide bandwidth. Fig. 4.1 shows the analysis model of a 30 mm monopole (\(\lambda/4\) at 2.5 GHz) and its spherical modal coefficients in 1–10 GHz range. The SME analysis is the decomposition of the radiated field from an antenna into orthogonal spherical modes and is a useful tool to understand antenna’s radiation properties [45]. At the low end (1 GHz) the mode spectrum is mainly composed of 48% TM\(_{01}\), 35% TM\(_{02}\), 7% TM\(_{04}\), 2.5% TM\(_{06}\), etc. whereas a 60 mm dipole has 100% TM\(_{01}\). The higher order TM\(_{02n}\) modes originate from the ground plane that prohibits (effectively, cancels out) the backward radiation (\(z < 0\)). This can be better understood from Fig. 4.2. When TM\(_{01}\) and TM\(_{02}\) modes have the 90° phase difference, the lower half of TM\(_{01}\) radiation is canceled out by that of TM\(_{02}\) radiation while upper halves are added, so that monopole-like pattern is obtained. To perfectly eliminate the backward radiation, an infinite number of higher order modes is required. Fig. 4.3 shows the radiation patterns of monopole antenna from 2 GHz to 9 GHz. As shown, the monopole-like pattern is “squeezed” as frequency increases up to 6 GHz, then from 7 GHz, the maximum radiation occurs at \(\theta = 45^\circ\). This is because the TM\(_{03}\) mode becomes the most dominant mode as the length of the monopole becomes three quarters of wavelength (\(A_H = 3\lambda/4\)). Note that TM\(_{04}\) mode is also increased to cancel out the backward radiation from the TM\(_{03}\) mode.

![Fig. 4.1. (a) Analysis model of a wire monopole antenna and (b) its spherical modal coefficients.](image-url)
Fig. 4.2. Combination of TM_{01} and TM_{02} modes with 90° phase difference results in a monopole-like pattern.

Fig. 4.3. Radiation patterns of the monopole antenna in 2–9 GHz.

Fig. 4.4. (a) Analysis model of a monocone antenna and (b) its spherical modal coefficients.

Fig. 4.5. Radiation patterns of monocone antenna over 2–9 GHz.
Another linear antenna with a monopole-like pattern is a monocone shown in Fig. 4.4. The analysis model and spherical modal coefficients of a monocone are shown in Fig. 4.4(a) and (b), respectively. For monocone, TM\(_{03}\) mode is not significant over the entire range but TM\(_{05}\) and TM\(_{06}\) modes are increased around 7 GHz. As a result, the main beam from the monocone is “squeezed” as frequency increases though the maximum radiation is fixed at the horizon as shown in Fig. 4.5.

![Figure 4.6](image)

**Fig. 4.6.** Geometry of the modified monocone (conical monopole) antenna.

![Figure 4.7](image)

**Fig. 4.7.** Simulated realized gain at horizon and minimum gain in FOV of the modified monocone for various conical junction height \(A_{H1}\).

Based on these radiation properties of monopole and monocone, a 90° monocone is modified to a conical monopole as shown in Fig. 4.6. Since the antenna is fed by a 50 Ω coax, the conical flare angle is 90° that gives approximately 50 Ω characteristic impedance for the conical transmission line region according to (2.12) and (2.13). The overall antenna height is 30 mm and the height of conical region (conical junction) \(A_{H1}\) is swept from 12.5 mm to 20.5 mm. For a short \(A_{H1}\), the antenna resembles a straight monopole, and so it does a full monocone when \(A_{H1}\) is long.
Fig. 4.8. Relative magnitudes of spherical modal coefficients of the modified monocone antenna. (a) TM$_{01,02}$ modes. (b) TM$_{03,04}$ modes. (c) TM$_{05,06}$ modes.
Fig. 4.7 shows the horizon gain and the minimum gain in the FOV for various $A_{HI}$, and Fig. 4.8 shows the spherical modal coefficients of TM$_{01-06}$ modes to analyze the gain variations. At 1 GHz, the mode spectrum is the same as that of a monopole/monocone. As frequency increases, TM$_{01}$ mode decreases and higher order modes increase, and the radiation pattern is distorted as a consequence. When $A_{HI} = 12.5$ mm, TM$_{03}$ and TM$_{04}$ modes are increased around 6 GHz and 4 GHz, respectively, due to the three quarterwave resonance of the monopole mode; as a result, the minimum gain in FOV is decreased around 6 GHz. TM$_{03}$ and TM$_{04}$ modes can be suppressed by increasing $A_{HI}$; however, TM$_{05}$ and TM$_{06}$ modes are increased above 6 GHz due to the contribution from the monocone mode, which is attributed to the degraded minimum gain in FOV at higher frequencies. To obtain widest 2 dBi bandwidth, $A_{HI}$ is chosen to be 16.5 mm that balances out TM$_{03,04}$ modes and TM$_{05,06}$ modes. As shown in Fig. 4.7, the 2 dBi bandwidth of the modified monocone antenna ranges from 1.75 GHz to 8.98 GHz (5.13:1) while the VSWR is below 2:1 above 1.66 GHz. Fig. 4.9 shows the simulated E-plane radiation patterns of the designed antenna on an infinite ground plane. As seen, excellent consistency of the monopole-like patterns over a wide bandwidth is achieved. Consistent patterns and the small diameter make this antenna a good candidate for the feeder of a wideband small reflector.

![Fig. 4.9. Simulated radiation pattern in E-plane of the modified monocone antenna with $A_{HI} = 16.5$ mm.](image)

**4. 3 Steerable Paraboloid Reflector**

A mechanically steerable reflector with the wideband modified monocone feeder analyzed in the previous section is discussed here.
4.3.1 Paraboloid Reflector

A half-cut dish-like paraboloid reflector on a ground plane is compared with a flat reflector over a wide frequency range. Fig. 4.10(a) and (b) show a flat reflector antenna with the monocone feed and gain response for various focal distances \( F_0 \), respectively. As expected, the gain bandwidth of the flat reflector antenna is limited by the half-wavelength destructive interference, i.e. when \( F_0 = \lambda/2 \), a critical gain drop arises since the reflected and direct waves are out-of-phase [85]. Fig. 4.11(a) and (b) show a paraboloid reflector antenna and the gain response for various focal distances \( F_0 \), respectively. The gain increases with frequency as the electrical size of the reflector increases, and the detrimental gain drops are much mitigated for the paraboloid. This is because the paraboloid is designed such that the propagation lengths of reflected...
rays are all equal at any incident angles \([103, 104]\), so that the reflected rays coherently interfere with the direct radiation from the feeder provided that the focal point of reflector coincides with the phase center of the feeder. Note that the phase center of the monocone is fixed at the origin (feeding point) throughout the entire bandwidth unlike a wideband monopole in Chapter 3. Fig. 4.11(b) also depicts the gain of ideal paraboloid reflector with aperture efficacy \(e_A = 100\%\) calculated using (4.1) \([105, 106]\).

\[
G = \frac{4\pi A}{\lambda^2} e_A \times 2 \tag{4.1}
\]

where the area of reflector aperture \(A\) is calculated as

\[
A = \frac{\pi R_w^2}{4} \tag{4.2}
\]
and $R_w$ is 120 mm herein. As shown, gain curves of the paraboloid follow the trajectory of the theoretical directivity with aperture efficiency for $F_0 = 42$ mm being better than 25% in the entire range.

![Fig. 4.12. Gain of paraboloid reflector with regular monocone feed for various $F_0$.](image)

Fig. 4.12 shows the gain of the paraboloid reflector for various $F_0$ fed by a regular 90° monocone with the same height. The deeper gain drops are observed around 2 GHz and 6–10 GHz. At the low frequency band, the close proximity of the feeder to the reflector has a detrimental effect on the gain due to the blockage. At 6–10 GHz band, the squeezed radiation from the monocone shown in Fig. 4.5 has negative impact on the radiation pattern. For the modified monocone feed, since the feed has a smaller diameter and consistent monopole-like patterns, stable far-field performance is achieved though the electrical length of the focal distance is short. In addition, VSWR of the paraboloid fed by the modified and the full monocones are shown in Fig. 4.13(a) and (b), respectively. For the modified monocone feed, good impedance match of VSWR $\leq 2$ is maintained up to 10 GHz after the antenna is turned on, and the turn-on frequency (where VSWR = 2) is a function of $F_0$. The turn-on frequency is increased for short $F_0$ and vice versa. Note that the turn-on frequency of the modified monocone without the reflector is 1.66 GHz. The degraded impedance match at low frequencies is due to the strong interaction between the feeder and reflector: an inductive coupling through currents on the ground plane and a capacitive coupling through electric field between the two. When the electrical distance is increased, these couplings are reduced. For the regular
monocone feed, the capacitive coupling becomes much stronger due to the closer proximity; consequently, the impedance match is more deteriorated. As seen in the above study, the clear advantage of using the modified monocone over a regular monocone as the feeder of electrically small reflector is shown.

Fig. 4.13. VSWR of the paraboloid reflector for various focal distances $F_0$ with (a) modified and (b) regular monocone feed.
4. 3. 2 Parametric Study

An extensive parametric study of design parameters of the reflector is performed to achieve improved performance with miniaturized electrical size. The reflector has three design parameters: focal distance $F_0$, the subtended angle in E-plane $A_E$, and the subtended angle in H-plane $A_H$. For fixed $A_E$ and $A_H$, the height $R_H$ and the width $R_W$ of the reflector are given as
Fig. 4.15. Parametric study of subtended angles in E- and H-planes of the paraboloid reflector represented by contour plots of (a) turn-on frequency, (b) in-band maximum VSWR, (c) electrical height, (d) electrical width /2 and (e) minimum horizon gain, (f) minimum FBR, and (g) ground delay variation. (h) Overlapped plot as a synthesis of design goals denoted as contour curves with arrows.
\[ R_H = f_0 \cdot \frac{\sin(A_E)}{\cos^2\left(A_E/2\right)}, \]  
(4.3)

\[ R_W = 2 \cdot f_0 \cdot \frac{\sin(A_H)}{\cos^2\left(A_H/2\right)}, \]  
(4.4)

Also, for fixed \( R_H \) and \( R_W \), \( A_E \) and \( A_H \) are calculated as

\[ A_E = 2 \cdot \tan^{-1}\left( \frac{R_H}{2 \cdot f_0} \right), \]  
(4.5)

\[ A_H = 2 \cdot \tan^{-1}\left( \frac{R_W}{4 \cdot f_0} \right). \]  
(4.6)

In order to find out the focal distance that allows for the smallest electrical size of the reflector, a parametric study of \( F_0, A_E, \) and \( A_H \) is performed. The electrical height and width of the reflector are measured at the turn-on frequency as \( R_H/\lambda_L \) and \( R_W/\lambda_L \) where \( \lambda_L \) is the wavelength at the turn-on frequency as shown in Fig. 4.14. It is observed that the minimum \( R_H/\lambda_L \) and \( R_W/\lambda_L \) are achieved for \( F_0 \) between 41mm and 43mm for the most combinations of \( A_E \) and \( A_H \). Therefore, the focal distance \( F_0 = 42 \) mm is chosen and the remaining two design variables are determined as follows.

Fig. 4.15(a-g) show the contour plots of turn-on frequency, in-band maximum VSWR after turn-on, electrical height and half-electrical width at the turn-on, minimum horizon gain, minimum FBR, and group delay variation within 1.5–10 GHz for various \( A_E \) and \( A_H \) of the paraboloid while the focal distance is fixed at 42 mm. As seen, the turn-on frequency can be decreased by increasing \( A_H \). However, electrical width and in-band maximum VSWR are increased. The minimum in-band mismatch loss is desired as discussed in Section 3.7.1. Since one critical goal of this work is to miniaturize the electrical size of reflector among
others, the tradeoff between the turn-on frequency and electrical width is considered. In this regard, subtended angle in E-plane needs to be minimized for small electrical height under the condition that antenna performance is not much compromised. Also, high FBR and small group delay variation are pursued for transient HPEM systems. Based on these contour plots, the range of subtended angles in E- and H-plane can be found by overlapping contour curves of design goals, which include: turn-on frequency < 1.66 GHz, in-band VSWR < 1.75, $R_H < 0.4 \lambda_0$, $R_W < 1.0 \lambda_0$, gain > 11.5 dBi, FBR > 19 dB, and group delay variation < 0.3 ns. The shaded area in Fig. 4.15(h) indicates the design parameters that meet all design goals. Based on this, we choose the design parameters of the paraboloid as: $F_0 = 42$ mm, $A_E = 79^\circ$, and $A_H = 93^\circ$, and the size of the reflector is 177.0 mm $\times$ 69.2 mm $\times$ 46.6 mm as shown in Fig. 4.16. Note this representation also indicates the sensitivity of the developed design on the small variations in the structural parameters.

### 4.3.3 Fabrication of Mechanical Scanning System

Because the feeder has omnidirectional pattern and the reflector is separated from the feeder, mechanical scanning of the beam can be performed by rotating the reflector while the feeder is fixed. Fig. 4.17(a) shows the first step to enable such a mechanical scanning reflector system. A metallic plate of 304.8 mm (12in) diameter, called GP1, is assembled with an antenna mount for the measurement purpose. To reduce the friction between the fixed and rotating metal plates, about five hundred steel balls with 1.57 mm (16 mil) diameter are laid in ball ditches. Also, lubricant oil is used to further reduce the friction. A motor is attached to the bottom of GP1, while the gear is protruded on top to be interlocked with the rotating plate. The picture of assembled GP1 is shown in Fig. 4.17(b) and (c).

The second step is illustrated in Fig. 4.18(a). On top of the fixed GP1, the rotatable metallic plate with diameter of 203.2 mm (8in) is placed and the gear is interlocked with the motor as shown in Fig. 4.18(b). Then GP2 is sandwiched between GP1 and GP3, and GP1, GP3 and the N-connector are assembled using screws. Therefore, GP2 can be rotated, while GP1, GP3, and the connector are fixed. Also, the end of
connector launch pin is threaded to be assembled with the monocone securely. The picture is provided in Fig. 4.18(c).

Fig. 4.17. Fabrication of steerable reflector antenna system: Step 1. (a) Configuration. (b) Overview picture. (c) Side view picture.
Fig. 4.19(a) shows the assembly steps of the antenna system. 304.8 mm (12in) diameter GP4 with 76.2 mm (3in) hole in the middle, which is the diameter of GP3, is put atop of GP2. Then GP3 and GP4 form a flat ground plane for the antenna. GP4 can rotate with the reflector, while GP3 is fixed with the feeder. The reflector, GP4, and GP2 are assembled as one piece by screws as shown so that they can rotate together. Also, the threaded launch pin of connector screws into the monocone. Finally, the motor is connected to the controller as shown in Fig. 4.19(b).
4. 3. 4 Measured Results

Fig. 4.20 shows VSWR and the realized gain of the proposed reflector antenna. An excellent agreement between simulations and measurements is obtained. The measured VSWR is below 2:1 from 1.66 GHz to
The measurement is performed up to 11 GHz, which is the cutoff frequency of used N-connector; however, the simulation data shows that the antenna is matched through 20 GHz and above. The upper frequency limit of monocone antenna is determined by the feeding coax and the gap and it can be increased by reducing the connector dimensions; however, the power handling capability of antenna will be reduced as a tradeoff. The size of the reflector at the turn-on frequency is 69.24 mm (0.383\(\lambda_0\)) \(\times\) 177.04 mm (0.980\(\lambda_0\)) \(\times\) 58.50 mm (0.324\(\lambda_0\)). The measured gain of the antenna is higher than 10 dBi from 2.15 GHz and no critical gain drop occurs owing to the small diameter and consistent radiation of the feeder.
Fig. 4.22. Measured E-plane pattern.
Fig. 4.23. Measured H-plane pattern.
Fig. 4.21 shows the FBR and elevation angle of the antenna. The elevation angle is direction of the maximum radiation with regard to the horizon in E-plane and FBR is defined in H-plane (xy-plane). The measured FBR is higher than 20 dB from 2.05 GHz up to 11 GHz, while the simulation shows that FBR is decreased at higher frequency band due to the diffracted field at the edges of the reflector. Also, the elevation angle is related to the finite electrical size of the ground plane. The measured E- and H-planes radiation patterns are shown in Fig. 4.22 and 4.23, respectively. As shown, good end-fire directional radiation patterns with high FBR are obtained over a wide bandwidth. Relatively high side lobes are due to the blockage from the feeder, which is a compromise for small reflector antennas, and diffraction at the edges of the reflector.

![Group Delay Graph](image)

Fig. 4.24. Simulated group delay with and without the reflector.

### 4.3.5 Time and Power Domain Performance

The antenna performances in time and power domains are investigated to assess possible use for transient HPEM applications. Fig. 4.24 shows the simulated group delay of the antenna with and without the reflector on an infinite ground plane. As expected, the group delay without the reflector decreases as frequency increases because the radiating region is closer to the feeding. With the reflector, group delay is increased due to the longer path of the wave propagation and oscillation is seen as a result of ringing.
between the feeder and the reflector; however, group delay variation is still less than 0.3 ns over the matched bandwidth. Using the second order Gaussian pulse with 50 ps pulse width as an input, the output pulses at the horizon with and without the reflector are shown in Fig. 4.25. Without the reflector, output pulse shape is almost the duplication of the input pulse with an increased magnitude. With the reflector, the same output pulse from the direct wave is followed by much stronger reflected pulse with ringing effect. The computed fidelity factors for cases without and with reflector are 96.4% and 89.1%, respectively. It is important to notice that ringing is generally not desired for HPEM applications and some efforts will need to be made.
in the future to mitigate them. The power spectral density (PSD) of input and output pulses are shown in Fig. 4.26. The 10 dB bandwidth of input pulse is from 0.90 to 9.98 GHz. The output pulse PSD without the reflector is almost overlapped with that of input, while that with the reflector is slightly shifted.

![Graph showing power spectral density](image)

**Fig. 4.27.** Maximum electric field strength and peak power handling capability.

![Electric field distribution on the antenna at 4 GHz](image)

**Fig. 4.28.** Electric field distribution on the antenna at 4 GHz.

Fig. 4.27 depicts the estimated maximum electric field intensity developed on the antenna and the air coax transition regions, and the corresponding peak power handling capability. The peak power handling capability $P_{\text{peak}}$ is calculated as

$$P_{\text{peak}} = \frac{E_c^2}{E_{\text{max}}/1W}$$  \hspace{1cm} (4.7)

where $E_c$ is the dielectric strength of the surrounding medium, which is herein air ($E_c = 3$ MV/m) and $E_{\text{max}}/1W$ is the maximum electric field intensity caused by 1 W input. The maximum electric field in the matched bandwidth ranges over 6–11 kV/m and peak power handling capability is between 0.1 MW to 0.25
Since the antenna is all metal, average power handling limitation due to the heat is not a major concern.

To better examine the critical region, the electric field distribution on the antenna is shown in Fig. 4.28. As shown, the strongest electric field is developed on the feeding coax. The theoretical peak power handling capability of the used coax transition is calculated as

\[
P_{\text{peak, coax}} = \frac{\pi E^2 a^2 \ln\left(\frac{b}{a}\right) \sqrt{\varepsilon_r}}{\eta_0} = 0.14 \text{MW}
\]

where \(a\) and \(b\) are inner and outer radii, \(\varepsilon_r\) is the permittivity, and \(\eta_0\) is the intrinsic impedance of the coax.

The correlation of the calculated value in (4.8) with the simulated peak power handling capability value confirms that the power handling is limited by the feeding coax. This indicates that the power handling can be improved by utilizing a customized coax with larger diameters and/or filling with pressurized gas [50].

4.4 Miniaturization of Monocone

Reducing the electrical size of the omnidirectional modified monocone antenna by engineering a parallel resonance is proposed in this section. Fig. 4.29 shows the equivalent circuit model of the modified monocone shown earlier in Fig. 4.6. This model is based on the circuit model of a monocone in Fig. 2.1. The conical and cylindrical regions are modeled as two transmission lines in series and are terminated by a series RLC circuit. Detailed interpretation of circuit model can be found in Chapter 2. The full-wave and circuit simulated impedances of the monocone are depicted in Fig. 4.30. A series resonance formed by \(L_A\) and \(C_A\) occurs at 1.42 GHz and stable \(50 + j0 \ \Omega\) impedance is maintained above.

For shorted top hat loaded monocone antennas in [98-101], the wire loading can be modeled as a tank LC connected in parallel with \(C_A\) and \(R_r\). Then \(C_A\) from the monocone and the inductance from the wire form a parallel resonance in addition to the inherent series resonance of the monocone. Since the inductance
of the wire is set due to the fixed wire dimension, the antenna capacitance $C_A$ controls the parallel resonant frequency. The separation between the parallel and series resonances needs to be wide enough to achieve good impedance match at low frequencies, which requires a large diameter of the hat, otherwise the impedance match is severely deteriorated. Furthermore, the strong self-resonance of wire often limits the inherent wide bandwidth of a monocone from bandwidth ratio $(br) > 10$ to $br \sim 3$.

Fig. 4.30. Full-wave and circuit simulated impedance of the monocone in Fig. 4.6.

Fig. 4.31. Modified monocone loaded with shorted semi-helical wires. (a) Configuration. (b) Equivalent circuit model.
4. 4. 1 Semi-Helical Loading

In order to control the parallel resonant frequency to reduce the turn-on frequency of antenna without increasing its diameter, the shunt inductance can be controlled by increasing the length of wire. This can be accomplished by replacing the straight wires with a semi-helical wires as shown in Fig. 4.31(a). Then the inductance of wire is determined by the helical turn ratio $H_T$. Moreover, the proposed wire loading can be integrated within the monocone volume, so that the physical size of the antenna remains the same. Fig. 4.31(b) shows the equivalent circuit model of the modified monocone loaded with two shorted semi-helical wires. The shorted wires attached at the conical junction are modeled as an inductor and a parasitic capacitor in parallel.

![Fig. 4.32. Full-wave and circuit simulated input impedance of the monocone antenna with and without the semi-helical loading. (a) Real parts. (b) Imaginary parts.](image)
The impedance and VSWR of the antenna with and without the load ing are shown in Fig. 4.32 and Fig. 4.33, respectively. Circuit-simulated impedances are also included and agree well with the full-wave simulations. The equivalent circuit parameters are listed in Table 4.1. Without the loading, the antenna has series resonance at 1.42 GHz and the turn-on frequency of VSWR = 2 is 1.66 GHz. The semi-helical loading generates parallel resonance below the unaltered series resonance, so that the input resistance is increased to reduce the turn-on frequency. The separation between series and parallel resonances and the turn-on frequency can be controlled by the helical turn ratio. For $H_T = 0.15$, 0.25, and 0.35, the turn-on frequency is 1.18 GHz, 1.26 GHz, and 1.34 GHz, respectively. Since lowering turn-on frequency increases in-band mismatch loss, $H_T = 0.25$ is chosen to trade off turn-on frequency against the in-band mismatch loss. Note that the 2 dBi bandwidth of the antenna with the proposed loading covers 1.22–7.33 GHz (6.01:1).

<table>
<thead>
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<th>0.15</th>
<th>0.25</th>
<th>0.35</th>
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<tr>
<td>$Z_{C1}$ ($\Omega$)</td>
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<td>49.61</td>
<td>49.42</td>
</tr>
<tr>
<td>$l_{m1}$ (mm)</td>
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<td>22.49</td>
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<td>67.00</td>
</tr>
<tr>
<td>$l_{m2}$ (mm)</td>
<td>23.51</td>
<td>23.07</td>
<td>24.44</td>
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</tr>
<tr>
<td>$C_A$ (fF)</td>
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<td>820.6</td>
<td>617.5</td>
</tr>
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<td>$C_H$ (fF)</td>
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<td>535.1</td>
<td>569.8</td>
</tr>
</tbody>
</table>

Table 4.1. Values for equivalent circuit parameters in Fig. 4.31(b) for various helical turn ratio $H_T$.

Fig. 4.33. VSWR of the monocone antenna with and without the semi-helical loading.

The impedance and VSWR of the antenna with and without the loading are shown in Fig. 4.32 and Fig. 4.33, respectively. Circuit-simulated impedances are also included and agree well with the full-wave simulations. The equivalent circuit parameters are listed in Table 4.1. Without the loading, the antenna has series resonance at 1.42 GHz and the turn-on frequency of VSWR = 2 is 1.66 GHz. The semi-helical loading generates parallel resonance below the unaltered series resonance, so that the input resistance is increased to reduce the turn-on frequency. The separation between series and parallel resonances and the turn-on frequency can be controlled by the helical turn ratio. For $H_T = 0.15$, 0.25, and 0.35, the turn-on frequency is 1.18 GHz, 1.26 GHz, and 1.34 GHz, respectively. Since lowering turn-on frequency increases in-band mismatch loss, $H_T = 0.25$ is chosen to trade off turn-on frequency against the in-band mismatch loss. Note that the 2 dBi bandwidth of the antenna with the proposed loading covers 1.22–7.33 GHz (6.01:1).
4.4.2 Measurement

The modified monocone and the semi-helical wires are fabricated using a computer numerical control (CNC) machining and a direct metal laser sintering (DMLS) 3-D printing, respectively, and those are assembled with an 863 mm circular ground plane and an N-connector feed, which is rated up to 11 GHz, as shown in Fig. 4.34. The end of launch pin in N-connector is threaded to be assembled with the monocone securely. Fig. 4.35 shows the measured and simulated VSWR of the monocone antenna with and without loading. The antenna without the loading has VSWR below 2:1 from 1.73 GHz to 20 GHz, and the proposed loading reduces the turn-on frequency to 1.23 GHz while the good impedance match is maintained through higher frequencies. Also, the self-resonance of the wire loading occurs around 5 GHz, which further improves impedance match.

Fig. 4.34. Fabrication of the monocone antenna loaded with semi-helical wires. (a) Configuration. (b) Picture.
Fig. 4.35. VSWR of the proposed antenna.

Fig. 4.36. Realized gain and WoW in H-plane.

The gain and omnidirectionality of the antenna, plotted as wobble-of-the-wave (WoW), with and without the loading are shown in Fig. 4.36. A stable gain curve around 5 dBi and decent omnidirectionality are accomplished for both. At the self-resonance of the wire (5.2 GHz), the measured gain and WoW are increased to 6.5 dBi and 7.9 dB, respectively. This is because the two wire loading induces TM_{22} and TE_{22} modes where the fundamental TM_{01} and TM_{02} modes decrease as depicted in Fig. 4.37. The measured radiation patterns of the antenna with loading in E- and H-planes are shown in Fig. 4.38 and Fig. 4.39,
respectively. Consistent monopole-like omnidirectional patterns in E-plane and H-plane are attained, though H-plane pattern is slightly squeezed at 5 GHz.

![Fig. 4.37. Spherical modal coefficients of the proposed antenna with loading and surface current distribution at the self-resonant frequency of wire as an inset.](image)

![Fig. 4.38. Measured radiation patterns in E-plane.](image)
4. 4. 3 Time and Power Domain Analysis

Using the same second order Gaussian pulse from Section 4.3.5 as an input, the output pulses with and without the loading are shown in Fig. 4.40. In both cases, excellent correlations with the input are achieved, demonstrating the minimum distortion in TD. Fig. 4.41 depicts the maximum electric field intensity developed on the antenna and the corresponding peak power handling capability. Without the loading, the maximum electric field strength is stable around 6–8 kV/m and the peak power handling capability is higher than 100 kW. With the loading, the field strength is almost the same except at the self-resonance of the wire. As shown in the inset of Fig. 4.37, a strong electric field is formed on the wire so that the power handling capability is limited. Again, the peak power handling capability of this antenna can be increased utilizing a customized coax and filling with pressurized gas.

Fig. 4.39. Measured radiation patterns in H-plane.
4.4.4 Resistive Termination

The turn-on frequency of the antenna can be further reduced by adding resistors within the antenna structure. Fig. 4.42(a) and (b) show the monocone loaded with the terminated semi-helical loading and its equivalent circuit model, respectively. The semi-helical wires are terminated by lumped resistors modeled as $R_t/2$. The resistor value determines the $Q$-factor of the parallel resonances so that the impedance
bandwidth can be further enhanced. Fig. 4.43 and Fig. 4.44 show the simulated impedance and VSWR of the antenna for various resistor values. As shown, the steepness of the parallel resonance is eased and the impedance match at very low frequencies is improved by increasing the resistance. For $R_T = 50 \, \Omega$, the input impedance at the low end is $25 \, \Omega$, which provides $VSWR = 2$, and VSWR is maintained below $2:1$ over the entire bandwidth. Fig. 4.45 shows the realized horizon gain and minimum gain in FOV of the monocone loaded with terminated wires. At low frequencies, increasing $R_T$ gives rise to the decreased gain; however, at high frequencies, the gain is not much affected by the termination since the semi-helical wire functions as an RF choke. Therefore, it is obvious that there is a tradeoff between the impedance match and the gain at low frequency band. Despite the small gain, the low VSWR can be preferred for transient HPEM antennas because the low frequency components of the high power pulse, though small, can damage the system if totally reflected. With this termination, the low frequency components of the input are dissipated by the resistor.
Fig. 4.43. Simulated input impedance of the monocone antenna loaded with the terminated semi-helical loading. (a) Real parts. (b) Imaginary parts.

Fig. 4.44. VSWR of the monocone antenna loaded with the terminated semi-helical loading.
4. 5 Conclusion

This chapter dealt with a challenging antenna problem in HPEM with practical applications – conceptual development of an antenna system that can easily be reconfigured to be directional (with scanned beam) and omnidirectional over a very wide bandwidth. To this end, a monocone antenna with and without a paraboloid reflector for steerable transient HPEM radiation is proposed and studied theoretically and experimentally. A modified monocone with wideband monopole-like omnidirectional pattern is used as the feeder, and its design process is analyzed using SME technique. The multi-dimensional parametric study of paraboloid reflector is conducted to miniaturize its electrical size while ensuring good frequency and time domain performances are maintained. A mechanical scanning system is prototyped to demonstrate the concept. The simulated and measured data show that the antenna has VSWR ≤ 2, gain > 10 dBi, FBR > 15 dB, and directional (to horizon) radiation patterns from 1.66 GHz to 20 GHz. The electrical size of the antenna at the turn-on frequency is only $0.383\lambda_0 \times 0.980\lambda_0 \times 0.324\lambda_0$. It is also shown that the proposed antenna can radiate short transient pulses with high peak power. Additionally, the semi-helical wire loading for the miniaturization of wideband omnidirectional monocone antenna is proposed. This antenna is also fabricated and measured, showing VSWR < 2 from 1.23 GHz, which is reduced by 33.8%, and consistent...
gain and monopole-like patterns up to 11 GHz. The proposed loading can be integrated with the reflector to achieve even smaller size of the reflector antenna, which will be performed in the future.
CHAPTER 5

Bi-/Uni-Directional Log-Periodic Antennas for High Power CW Radiation

5.1 Introduction

Log-periodic (LP) antennas were introduced in 1958 [107] and since then they have been used in a variety of applications such as military and commercial communications, EMI/EMC testing, radar, etc. from HF to optical frequencies [108-111]. LP antenna is essentially a series of carefully arranged dipoles fed through a single feed line. By having a certain scale rate between consecutive dipoles, the impedance and far-field properties may repeat logarithmically over a multiple octave bandwidth. While LP dipole linear array has end-fire radiation patterns, a planar embodiment of LP antenna on a substrate has bi-directional pattern that is useful for relay and MIMO systems [112]. To reduce complexity and cost while keeping the inherent bi-directionality, a microstrip feed is preferred over a coaxial excitation. This feed can also be easily designed to act as an impedance transformer thus improving match from an inherently high impedance of LP antennas (~188.5 Ω for self-complementary aperture in free space) to 50 Ω [113]. Since the feeding line of an LP aperture, often called boom, functions as a ground plane for the microstrip line, the flare angle of the boom needs to be large enough to provide wide enough ground plane as in [114]; otherwise, guided-wave in the microstrip line couples to radiating dipole elements, deteriorating the quality of the radiation and impedance match. However, the wide boom angle comes with sacrifice in the turn-on frequency due to the shortened dipole elements [115].

Even if a planar LP antenna is well-designed in frequency domain (FD), it can be susceptible to microwave breakdown in power domain (PD). For a high peak short pulse input, the major concern is the dielectric breakdown, whereas the thermal breakdown is more critical for continuous wave (CW) and damped sinusoidal signal input [50]. Even for a short pulse input, thermal effects may take place when the pulse repetition frequency (PRF) is high, leading to a high average power $P_{av}$ computed as

$$P_{av} \approx P_{peak} \cdot \text{FWHM} \cdot \text{PRF}$$  (5.1)
where $P_{\text{peak}}$ is the peak power and FWHM stands for full width at half maximum. Such thermal failure can cause expansion and detachment of traces, change in the material property, melting dielectric, structural deformation, just to mention a few. In spite of possible hazards, the thermal performance of antennas and other RF devices has been seldom considered.

In this chapter, a planar bi-directional LP antenna is designed and its performance is analyzed in electromagnetics (FD) and thermal domain (ThD). Novel wide-boom geometry is proposed and the improved match and far-field characteristics are demonstrated with an LP antenna fed by a microstrip impedance transformer without sacrificing its turn-on frequency. Electro-thermal multiphysics analysis is performed to evaluate the thermal behavior of the antenna under high power CW input. It is seen that the applied geometrical modification improves antenna performance not only in FD, but also in ThD. Measurements verify that the proposed antenna has good impedance match and flat gain over a broad bandwidth. Also, the temperature distribution on the antenna is measured using non-contact thermal imaging system, and the good correlation with multiphysics simulations is used to showcase the robustness of the developed model.

For long distance point-to-point communication systems or radars, broadband uni-directional antennas with high power handling capability are required. Ridged waveguide horn antenna is the typical choice for such applications. However, horn antennas are inherently bulky, heavy, and costly [116-119]. For example, the horn antenna in [116], which works from 1 GHz to 8 GHz, has the dimensions of 235 mm ($0.78\lambda_0$) × 252 mm ($0.84\lambda_0$) × 175 mm ($0.58\lambda_0$) and weighs 1.4 kg while power handling capability is 200 W for CW input. In this chapter, as an alternative to horn antennas, a wideband flush-mountable uni-directional LP antenna with smaller size, lighter weight, and comparable performance is presented.

Since planar embodiments of LP antennas are bidirectional, a cavity backing on one side of the antenna aperture is typically used to achieve uni-directional patterns and the desired mount. A variety of cavity backing strategies has been explored in the past. From the ray perspective, it is argued that an empty metallic cavity underneath the radiator reflects backward radiation to the forward direction and is most effective when the depth is a quarter-wavelength [120-122]. However, this reflective cavity backing limits antenna
bandwidth at both high and low frequency bands. At the high frequency band, the reflected wave destructively interferes with the direct wave, resulting in distorted radiation patterns. To avoid this interaction, the depth of cavity should be limited to less than a half-wavelength at the highest operating frequency. At the low band where the electrical distance between the antenna and reflector is short, the impedance match is deteriorated due to the disturbed current on the antenna.

To extend the bandwidth of cavity-backed antennas, the shape of a cavity is modified in [123-125]. A stepped reflector [125] provides conceptually a constant electrical depth for each radiating element to avoid the destructive interference over a broad bandwidth. However, this causes multiple reflections inside the cavity, beam broadening, and increased back lobes. Another technique used to keep the electrical depth constant is the layered dielectric loading [121]; however, this approach is not sufficiently wideband and some issues with resonances may arise.

Both good impedance match and gain over wide bandwidth can be achieved by absorber-loaded cavity backing that dissipates the backward radiation [120, 126, 127]. However, absorptive backing inherently reduces the efficiency by more than 50% throughout the bandwidth. To overcome this, cavities partially loaded with absorber are reported. Various absorber deployments have been suggested in the literature to maximize the gain and minimize VSWR [128-130]. These methods require precisely engineered absorber block (for example ferrite), which is often difficult to fabricate.

A novel wideband cavity backing is proposed in this chapter. A bowtie slot is etched in the cavity bottom to reduce the interaction with the antenna aperture, thus improving the impedance match. The backward radiation from the slot is suppressed by a microstrip termination and an additional cavity on the bottom. In addition, it is conceptually shown that the microstrip termination of the slot enables dissipated power in the slot to be recycled. The proposed cavity-backed flush mountable uni-directional LP antenna achieves VSWR ≤ 2.5, gain > 0 dBi, and front-to-back ratio (FBR) > 7 dB from 0.59 GHz to 5 GHz.

This chapter is organized as follows:

- Section 5.2 proposes a wide-boom geometry bi-directional LP antenna.
- Section 5.3 describes modeling of antennas in electro-thermal domain.
- Section 5.4 characterizes the bi-directional LP antenna in electro-thermal domain.
- Section 5.5 designs a cavity backing for a uni-directional coverage of LP antenna.
- Section 5.6 concludes the chapter.

5.2 Wide-Boom Bi-Directional Log-Periodic Aperture

The wide-boom geometry of LP antenna is proposed and its wideband bi-directional performance is analyzed and measured in FD.

![Geometry of the proposed wide-boom LP antenna. Detailed view of the boom is shown and compared with the conventional boom geometry.](image)

5.2.1 Antenna Design

Fig. 5.1 shows configuration of the proposed wide-boom LP antenna implemented on a 1.524 mm RO4003 substrate ($\varepsilon_r = 3.55$, $\tan\delta = 0.0027$). Its design parameters are summarized in Table 5.1. Planar LP
The feed is connected to the aperture through a plated via. A dummy microstrip line is included for pattern symmetry. Due to the presence of dielectric, the nominal impedance of LP aperture is \( \sim 140 \, \Omega \); thus, the microstrip line is linearly tapered from 3.33 mm (50 \( \Omega \)) to 0.4 mm (122.9 \( \Omega \)). The boom of an arm provides ground plane for the microstrip line, where the flare angle \( \beta \) is fixed at 10\(^\circ\) to keep the turn-on frequency low. However, the
small boom angle renders the ground plane narrow, especially around the center; consequently, the characteristic impedance of the feed line is higher than it is designed for. To provide a wide ground plane with small boom angle, virtual axis (apex) of the fan-shaped boom is offset by $B = 40$ mm. The magnified view of the proposed boom illustrates that the ground plane becomes much wider than the conventional boom.

The procedure to widen the boom is illustrated in Fig. 5.2. Starting from a conventional LP aperture shown in Fig. 5.2(a), the boom axis is moved by $B = 40$ mm and the boom is cut at the bow-tie feed region,
making the boom wider as depicted in Fig. 5.2(b). Note that the stability of the patterns and impedance over broad bandwidth is compromised with this geometry since the aperture is not self-complementary. To maintain self-complementarity, the booms are rotated by 90° with duplication in Fig. 5.2(c), and duplicated booms are subtracted from LP dipole elements in Fig. 5.2(d). A minimum width of the boom is increased to 8.25 mm (from 1.44 mm).

Figs. 5.3(a) and (b) show VSWR and gain of LP antennas for various offsets of boom apex \( B \). A conventional boom in Fig. 5.2(a) and the proposed wide-boom in Fig. 5.2(d) are denoted as \( B = 0 \) mm and 40 mm, respectively. These and all other RF simulations are performed in Ansys HFSS. For \( B = 0 \) mm, multiple peaks in VSWR and gain drops are caused by the narrow ground plane, whereas those are much improved for \( B = 40 \) mm. Note that the turn-on frequency of VSWR ≤ 2 is about the same. Two-port analysis setup depicted in Fig. 5.4 is used to understand these observations. The LP aperture is cut in half and two ports are assigned to the ends of microstrip impedance transformer. This feed line is also simulated on an infinite ground plane as a benchmark. Fig. 5.5 shows the time domain reflectometry (TDR) impedance profile of the feed line on the half-cut LP aperture with various offsets of the boom and on an infinite ground plane. 0 ns and 0.75 ns represent ports 1 and 2, respectively. The impedance on an infinite ground plane is 50 Ω at port 1 and increases to 127 Ω at port 2. Ideally, the TDR impedance on the aperture will be identical to that on an infinite GP. However, when \( B = 0 \) mm, the impedance reaches 139 Ω due to the narrow boom around port 2. When \( B = 40 \) mm, TDR impedance profile is closer to the benchmark since the boom is
wider. This results in stabilized feeder impedance and improved impedance match of the antenna as shown in Fig. 5.3(a).

Fig. 5.5. TDR impedance profile of the fed line on the half-cut LP aperture for various boom offsets (Fig. 5.5) and on an infinite ground plane.

Fig. 5.6. Insertion loss of the feed line from the two port analysis.

Fig. 5.6 shows the insertion loss of the feed line on the half-cut LP aperture and on an infinite ground plane. As seen, the insertion loss of the proposed boom correlates well with the benchmark, while that of the conventional boom is higher and has multiple dips. Fig. 5.7(a) and (b) show current densities on conventional and the proposed booms, respectively. As seen, the current on a conventional boom excites
radiating elements, leading to coupling/radiation loss. This coupling is much suppressed on the proposed wide-boom so the gain drops of the antenna are alleviated for the proposed geometry (Fig. 5.3(b)).

![Surface current distribution at 3.4 GHz](image)

Fig. 5.7. Surface current distribution at 3.4 GHz on (a) conventional ($B = 0$ mm) and (b) wide-boom ($B = 40$ mm) LP aperture induced by the microstrip feed.

![VSWR and picture of the proposed LP antenna](image)

Fig. 5.8. VSWR and picture of the proposed LP antenna.

### 5.2.2 Measurement

The proposed wide-boom LP antenna is manufactured on a 2oz copper plated 1.524 mm RO4003 dielectric using standard PCB process as shown in the inset of Fig. 5.8. Since thermal imaging of this antenna will be performed in the next section, black soldermask thinner than 25 μm covers the copper layer.
of antenna, which obscures thermal imaging due to low emissivity. The antenna is assembled with an epoxy glass holder, which is fabricated using laser cutting machine, to be installed on the rotator in an anechoic chamber as shown.

The measured and simulated VSWR of the antenna are shown in Fig. 5.8. The measured VSWR is below 2 from 0.43 GHz to 5 GHz. The electrical size of the antenna at 0.43 GHz is 0.358 λ₀ x 0.358 λ₀ x 0.02 λ₀. The simulated data agree very well with measurements. Fig. 5.9 shows the realized gain. The flat gain around 5 dBi and cross-pol rejection higher than 6.5 dB are achieved. The radiation patterns in E- and H-planes are shown in Fig. 5.10.

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H-planes are shown in Fig. 5.10(a) and (b), respectively. Stable bi-directional radiation patterns are confirmed over the entire bandwidth.

5.3 Electro-Thermal Modeling of Antennas

This section describes the theoretical basis of RF heat generation and distribution, and phenomenological analysis using electro-thermal multiphysics simulation. A microstrip patch antenna is modelled in the multiphysics domain as an example. A microstrip patch antenna is chosen because of its simplicity and well-understood operating principle. The temperature distribution of the patch antenna is explained based on the electromagnetic field distribution.

5.3.1 Theory

Thermal heating of an antenna is caused by RF losses including conduction, dielectric, and magnetic losses as derived from the power loss term in Poynting’s theorem (5.2) [131].

\[
P_l = \frac{\sigma}{2} \int |E|^2 \, dv + \frac{\omega}{2} \int \left( \epsilon'' |E|^2 + \mu'' |H|^2 \right) \, dv. \tag{5.2}
\]

Since the used materials are non-magnetic, the magnetic loss can be neglected. The conduction loss \( P_{lc} \) due to the finite conductivity of a metallic strip is given as

\[
dP_{lc} = \frac{\sigma}{2} |E|^2 \, dv_c, \tag{5.3}
\]

and applying \( J_s = \sigma E \) gives

\[
dP_{lc} = \frac{1}{2\sigma} |J_s|^2 \, dv_c \tag{5.4}
\]

where \( \sigma \) denotes the electric conductivity, \( |E| \) is electric field intensity, \( |J_s| \) is surface current density, and \( dv_c \) is the infinitesimal volume of a conductor. Since the conduction loss is more associated with current density and is inversely proportional to the electric conductivity, expressing the conduction loss as (5.4) is more reasonable than (5.3). In addition, in a good conductor (\( \sigma \gg \omega \epsilon \)), (5.4) can be rewritten as

\[
dP_{lc} = \frac{\delta}{2\sigma} |J_s|^2 \, ds \tag{5.5}
\]
where $\delta_s$ is the skin depth, given as $\sqrt{\frac{2}{\omega \mu \sigma}}$.

The second heating mechanism is the dielectric loss $P_{ld}$

$$dP_{ld} = \frac{\omega}{2} \varepsilon'' |E|^2 \, dv_d$$

where $\varepsilon''$ is imaginary part of permittivity and $dv_d$ is the infinitesimal volume of the dielectric. Thus the total (incremental) RF loss of the antenna is given as

$$dP_l = dP_{lc} + dP_{ld} = \frac{1}{2\sigma} |J_s|^2 \, dv_c + \frac{\omega}{2} \varepsilon'' |E|^2 \, dv_d.$$  \hspace{1cm} (5.7)

From (5.7), the RF heating per unit volume $Q$ can be written as

$$Q = \frac{1}{2\sigma} |J_s|^2 + \frac{\omega}{2} \varepsilon'' |E|^2.$$  \hspace{1cm} (5.8)

For electro-thermal simulations, this RF loss in the antenna is calculated in Ansys HFSS and mapped into Ansys Mechanics thermal solver using Ansys Workbench multiphysics platform as shown in Fig. 5.11.

Fig. 5.11. Multiphysics simulation setup in Ansys Workbench.

The heat energy developed on an antenna diffuses over the structure through conduction, convection, and radiation. The heat conduction $Q_c$ in a homogeneous medium can be expressed using the Fourier heat conduction equation [132],

$$Q_c = \rho C_p \frac{\partial T}{\partial t} - K \nabla^2 T$$  \hspace{1cm} (5.9)

where $\rho$ is density, $C_p$ is specific heat, $T$ denotes temperature, $t$ is time, and $K$ is thermal conductivity. In a steady-state condition, (5.9) is simplified to

$$Q_c = -K \nabla^2 T.$$  \hspace{1cm} (5.10)

In addition to the heat conduction through medium, overall cooling due to the natural air convection is
expected. Assuming the surrounding air is at a constant temperature $T_0$, the heat transfer $Q_s$ on surface $A$ can be calculated using Newton’s law of cooling,

$$Q_s = hA(T - T_0)$$  \hfill (5.11)

where $h$ is the heat convection coefficient. Another heat transfer mechanism is thermal radiation that is the loss of heat energy by electromagnetic radiation due to the motion of charged particles. The flowchart of electro-thermal multiphysics analysis is depicted in Fig. 5.12.

![Flowchart of electro-thermal multiphysics analysis.](image)

**5.3.2 Example: Microstrip Patch Antenna**

A simple microstrip patch antenna is analyzed in electro-thermal domain first. The patch antenna is chosen because the radiation mechanism is relatively simple and well understood. Fig. 5.13 shows the

![Microstrip patch antenna and its S11.](image)
layout of a microstrip patch antenna fed by a microstrip line through a quarter-wavelength impedance transformer on a 20 mil RO3003 ($\varepsilon_r = 3.0$, $\tan\delta = 0.001$) substrate and its S11 response. Figs. 5.14(a) and (b) show the electric field distribution in the dielectric and the current distribution on the conductor surface, respectively. As expected, the strongest electric field and current occur at the quarter-wavelength transformer due to the narrowest strip width. In the patch, the strongest electric field is formed at the edges where the radiation takes places. Also, the strongest current is formed in the middle of the patch where the electric field is minimal. Since the locations of strongest electric field and current are different, the impact of conduction and dielectric losses can be identified from the temperature distribution.

Fig. 5.14. Field distribution of the patch antenna at the resonant frequency. (a) Electric Field. (b) Current.

<table>
<thead>
<tr>
<th>Thermal conductivity $K$</th>
<th>Specific heat $C_p$</th>
<th>Thermal diffusivity $\alpha$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5 W/m·ºC</td>
<td>900 J/kg·ºC</td>
<td>2.65e-7 m²/s</td>
</tr>
</tbody>
</table>

Table 5.2. Thermal property of RO3003 dielectric substrate.

Let us first consider the dielectric loss given in (5.6) while the conduction loss is excluded. This can be done by assigning PEC ($\sigma = \infty$) boundary condition on conductors and finite tangent loss of $\tan\delta = 0.001$ in the substrate. Ambient temperature and the convection coefficient are assumed to be 20ºC and 10 W/m²ºC, respectively. Also, 100W CW power is used to feed the antenna. Since the temperature rise is linearly proportional to the input power, temperature of any other power level can be readily calculated. Table 5.2 lists the thermal property of the dielectric. The calculated RF heat generation is shown in Fig. 5.15. As seen, the most heat is generated at the edges of the patch as the strong electric field from the patch
passing through the lossy dielectric. The steady state temperature distribution due to the dielectric loss only is shown in Fig. 5.16. Hot spots (temperature ~70°C) are observed in the dielectric below the edges of the patch because of the strong electric field from the patch and poor thermal conductivity of the dielectric.

Fig. 5.15. RF heat generation due to the dielectric loss.

Fig. 5.16. Temperature distribution due to the dielectric loss only.

<table>
<thead>
<tr>
<th>Thermal conductivity $K$ (W/m·ºC)</th>
<th>Specific heat $C_p$ (J/kg·ºC)</th>
<th>Thermal diffusivity $\alpha$ (m²/s)</th>
<th>Density $\rho$ (kg/m³)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.5</td>
<td>900</td>
<td>2.65e-7</td>
<td>8930</td>
</tr>
</tbody>
</table>

Table 5.3. Thermal property of copper conductor.

Now let us consider the conduction loss given in (5.4) and (5.5). To exclude the dielectric loss, tangent loss of the substrate is assigned to be zero in HFSS simulation. The conduction loss can be assigned by two methods: finite conductivity boundary condition on a 2D flat surface and 3D solid conductor with finite
conductivity. In the former method, the current distribution is assumed to be uniform along the thickness as in (5.5) and the thickness is assumed to be equal to the skin depth. In the latter one, full integration over the volume of conductor is performed as in (5.4). Since the finite element method (FEM) meshing inside of the conductor in HFSS cannot accurately represent the current distribution, the latter method is not numerically dependable. Instead, finite conductivity on the finite thickness trace while omitting the mesh inside can account for the conduction loss on top, bottom, and side facets of the conductor, whereas the former method only accounts for the loss on one facet (bottom one in case of the patch). In this sense, we use the 3D solid conductor without meshing inside. A 1oz copper ($\sigma = 5 \times 10^8$ S/m) is used as the conductor for both strip and ground plane and its thermal property is shown in Table 5.3. Fig. 5.17 shows the temperature distribution of the patch antenna in which only the conduction loss is included. The peak temperature of 154ºC occurs in the dielectric below the center of the patch where the maximum current density is located.

With both dielectric and conduction losses, the temperature distribution is calculated in Ansys Mechanics as shown in Fig. 5.18. As seen, the maximum temperature occurred at the center of the patch, meaning that the conduction loss is more dominant than the dielectric loss for this patch antenna. The decrease in peak temperature is due to the increase in return loss and less accepted power. Since the thermal breakdown temperature of the RO3003 dielectric is 500ºC, the patch antenna is expected to handle 500W.
of CW input. In addition, note that the dominant loss factor can be changed depending on the material properties of dielectric and conductor. Fig. 5.19 shows the temperature distribution with lossier dielectric substrate but same conductor. In this case, the maximum temperature occurs at the edges of the patch.

Fig. 5.18. Temperature distribution due to both the conduction ($\sigma = 5 \times 10^8$ S/m) and dielectric ($\tan\delta = 0.001$) losses.

Fig. 5.19. Temperature distribution due to higher dielectric ($\tan\delta = 0.02$) losses.

5.4 Electro-Thermal Characterization of Log-Periodic Antenna

Based on the developed electro-thermal model of antenna in Section 5.3, the bi-directional LP antenna in 5.2 is analyzed in electro-thermal domain. Using the simulated electromagnetic loss data, the setup for thermal simulation is shown in Fig. 5.20. To obtain accurate result, the minimum and maximum lengths of
The tetrahedral mesh in the thermal simulation are set to be 0.02 mm and 5 mm, respectively. In addition, to accurately represent circular structure, the curvature angle of mesh is limited to be less than 15°. The natural convection coefficient is assumed to be 10 W/m²°C with 20°C ambient temperature and no other cooling mechanism is applied. Because the antenna will be fed through a coax cable in the actual experiment, an SMA connector and a 200 mm long coax cable are modeled, where an isothermal boundary condition is applied at the end. The thermal properties of the used substrate and metallic trace are summarized in Table 5.4 [133].

### Table 5.4: Thermal properties of used materials.

<table>
<thead>
<tr>
<th>Material</th>
<th>Rogers4003</th>
<th>Copper</th>
</tr>
</thead>
<tbody>
<tr>
<td>Conductivity (W/°K/m)</td>
<td>0.71</td>
<td>401.0</td>
</tr>
<tr>
<td>Specific Heat (J/°K/kg)</td>
<td>900</td>
<td>390</td>
</tr>
<tr>
<td>Diffusivity (m²/s)</td>
<td>4.41e-7</td>
<td>1.15e-4</td>
</tr>
<tr>
<td>Density (kg/m³)</td>
<td>1790</td>
<td>8930</td>
</tr>
<tr>
<td>Thickness (mm)</td>
<td>1.524</td>
<td>0.07</td>
</tr>
</tbody>
</table>

Fig. 5.20. Thermal simulation setup.

5.4.1 Multiphysics Analysis

Fig. 5.21 shows the simulated steady-state maximum temperature on the antenna for various $B$ when antenna is excited with 100 W CW signal at each frequency. For $B = 0$ mm, there are multiple large temperature spikes and the highest peak is 918°C at 3.34 GHz. Since the decomposition temperature of RO4003 is 425°C [133], the catastrophic effects are expected. Increasing $B$ significantly reduces the
temperature rise by decreasing RF losses as described in the previous section. In addition, wider boom provides larger area for heat conduction and convection that also help reduce the temperature rise. With the proposed wide-boom geometry, the highest temperature developed with 100 W is only 130ºC. Assuming the material properties remain unchanged with increased temperature, one can estimate power handling of ~ 370 W.

It is interesting to note there are multiple temperature peaks that follow the LP growth rate of the antenna as shown in Fig. 5.22–5.24. The maximum temperature is shown in the first row (replica of Fig. 5.21); the second row is the input reactance of the aperture (lumped port feed, i.e. no microstrip line); the radiation efficiency is in the third row; and the fourth row depicts the polarization squint with regard to x-axis. Vertical grid lines are drawn at the frequencies of local maxima in the temperature plot and horizontal grid lines are drawn at zero reactance. The uniform distribution of vertical grid lines on the log-scaled frequency axis indicates that the temperature peaks appear logarithmically. Also, local maxima in temperature correlate with resonant frequencies of the LP aperture. This is not surprising since at these frequencies the stronger fields are formed at the radiating elements leading to larger electromagnetic loss and higher temperature. Moreover, peak temperatures coincide with the local minima of radiation efficiency.
caused by the increased RF losses. Additional radiation efficiency drops are caused by the dummy microstrip line; however, the overall efficiency and pattern quality are improved.

The fourth row in Fig. 5.22–5.24 shows the azimuthal polarization squint of the LP antenna with regard to x-axis at boresight ($\theta = 0^\circ$). The polarization nature of a LP antenna is linear and is comparable to x-oriented dipole (see Fig. 5.1). However, the polarization axis of the LP antenna is horizontally squinted due to the geometry of the radiating dipoles. As the frequency changes from low to high, the active region of the antenna moves towards center, while alternating between left and right dipole elements. For example, at 0.49 GHz, the active dipole elements are the ones with radius $r_1$ on the second and the fourth quadrant, and the ones with radius $R_2$ at the first and the third quadrant will be the next active region at frequencies

![Graph showing maximum temperature, input reactance, radiation efficiency, and polarization squint in phi axis of LP antenna for $B = 0$ mm.](image)
slightly higher. This shift of the active region occurs logarithmically following the growth rate of LP aperture. When active dipole elements are in second and fourth quadrants, the polarization axis is squinted to the left with regard to x-axis, and vice versa. As seen, the horizontal polarization squint with regard to x-axis oscillates from negative to positive values with a uniform periodicity on the log-scale frequency axis. It is noticed that this logarithmic polarization squint also corresponds to the peak temperature frequencies and resonant frequencies. In summary, the stacked plots in Fig. 5.22–5.24 clearly demonstrate that at each resonance the LP aperture delivers accepted power to the active dipole with no excess stored energy.
Moreover, the polarization of the beam is aligned to the ends of active dipole where strong electric fields generate high loss and high temperature.

5.4.2 IR Measurement

Fig. 5.25 shows the thermal measurements setup. A signal generator is used to produce CW waveform with 10 dBm output power. A 30 dB gain power amplifier raises the level of the signal to ~40 dBm. A 30 dB directional coupler ensures that the small portion (0.1%) of the power is routed to the calibrated spectrum analyzer for monitoring output power, while the rest is carried to the antenna through a coaxial cable with
~5 dB insertion loss. Having measured the transmission coefficients of thru and coupled ports of the coupler as $S_{thru}$ and $S_{coupled}$, the input power to the antenna $P_{in}$ can be calculated as

$$P_{in} = P_{S_A} \cdot \frac{S_{thru}}{S_{coupled}} \cdot S_{coax}$$

(5.12)

where $P_{S_A}$ is the power reading from the spectrum analyzer and $S_{coax}$ is the transmission coefficient of the feeding cable.

To measure temperature distribution on the antenna under this input, a non-contact infrared (IR) thermal imaging camera is placed 1 m from the antenna as shown in the picture. FLIR’s A655sc uncooled micro-bolometer IR camera with spectral range of 7.5–14.0 μm and 640×480 resolution is used. The measurement is performed in anechoic chamber to avoid thermal noises such as human body, light bulb, electronics, etc. This way the noise sources in anechoic chamber are IR camera and its power supply.
Fig. 5.26. Steady-state maximum temperature rise.

Fig. 5.27. Field distribution of the antenna at resonance (1.79 GHz). (a) Electric field intensity. (b) Surface current density.
Fig. 5.26 shows the steady-state maximum temperature rise above the ambient and the calculated input power to antenna. Though the signal generator produces constant 10 dBm output from 1.71 GHz to 1.85 GHz with 0.01 GHz step, the actual input power level to the antenna varies because of the changing gain of power amplifier and cable loss. The antenna is left fed at each frequency until the steady-state has been reached (~ 6 min). Then, the power amplifier is turned off and natural air convection is used to cool down the antenna to the ambient temperature before the next test is conducted. This procedure is iterated at each frequency for front and back sides of the antenna. The temperature rise is taken by subtracting images of
ambient state (cooled down) antenna from recordings of radiating antenna in FLIR ResearchIR software to account for the change in room temperature over time. For multiphysics simulation, the same input is applied to the electro-thermal model of the antenna to be compared with the measurement. The convection coefficient is assumed to be 5 W/m²/ºC and the room temperature is set to 0ºC to see the temperature rise.

In Fig. 5.26, the measured maximum temperature rise is 13.7ºC at 1.77 GHz where tenth LP dipole element resonates, and the temperature rise curve is flat above 1.78 GHz. In simulations, the resonant frequency is found at 1.79 GHz. The shift in resonant frequency is due to the soldermask layer. In addition,
it is noticed that the temperature rise is higher for the back side measurement. Since dielectric loss is more dominant than conduction loss especially at resonant frequencies for the LP antenna (this may be reversed if dielectric loss tangent of the substrate is very small), the hot spots are located at which the strongest electric field is formed. Fig. 5.27(a) and (b) depict the simulated electric field and current distribution on the antenna at the resonant frequency of the tenth dipole element (1.79 GHz) to show where dielectric and conduction losses are strong. The electric field intensity is high at the end of active dipole element and the current intensity is strong at the junction between dipole and boom. Therefore, hot spots are expected to be

Fig. 5.30. Measured antenna temperature map at 1.85 GHz. (a) Top. (b) Bottom.
Located at the end of active dipoles. Also, the dielectric is expected to be hotter than copper layer because the produced heat is ‘trapped’ due to the low thermal conductivity. Therefore, the hot spot is formed below the end of active dipole. While the IR camera detects thermal radiation from objects to measure temperature, the thermal radiation from the dielectric will be reflected back to the bottom side (-z direction) by the copper layer. Therefore, it is difficult for camera to detect a true hot spot in the front side measurement.

Fig. 5.28(a) and (b) show the measured steady-state temperature map of the antenna at 1.77 GHz from top and bottom, respectively. The hot spot is found at the end of active (tenth) dipole as expected. Note that the lower arm has the higher temperature owing to the contribution from microstrip line. The simulated temperature map of the antenna at 1.79 GHz is shown in Fig. 5.29 and agrees well with Fig. 5.28; thus, the

Fig. 5.31. Simulated antenna temperature map at 1.85 GHz. (a) Top. (b) Bottom.
robustness of electro-thermal modeling of antenna is confirmed. Fig. 5.30 and 5.31 show the measured and simulated temperature map of the antenna, respectively, at 1.85 GHz, which is a non-resonant frequency. As seen, the highest temperature is much lower than at the resonant frequency.

![LP Aperture](image)

Fig. 5.32. LP antenna backed by an empty cavity.

![VSWR](image)

Fig. 5.33. VSWR of the LP antenna with empty cavity backing for various cavity height $H_l$.

5. 5 Cavity-Backed Uni-Directional Log-Periodic Antenna

The bi-directional LP antenna proposed in the previous section is transformed to uni-directional by adding a cavity backing. This section presents a novel slot-loaded cavity backing for wideband uni-directional LP antenna.
Fig. 5.32 shows the LP aperture backed by an empty cavity with height $H_1$, and its VSWR and realized boresight ($\theta = 0^\circ$) gain for various $H_1$ are shown in Fig. 5.33 and 5.34, respectively. When $H_1$ is short, the cavity deteriorates VSWR response due to the strong interaction with the aperture; specifically, the feeding current on the boom is disturbed when the electrical distance of $H_1$ is short. This undesired interaction can be reduced by increasing $H_1$; however, larger $H_1$ limits the gain bandwidth due to the destructive interference, i.e. reflected wave from the cavity has $180^\circ$ phase difference with the direct wave when $H_1 = \lambda/2$. The electrical distance $D_0$ between the aperture and the cavity back is calculated as

$$D_0 = \left( h_1 \cdot \sqrt{\varepsilon_{r1}} + h_2 \cdot \sqrt{\varepsilon_{r2}} \right) / \lambda_0$$

(5.13)

where $h_1 = h_2 = 60$ mil are heights of substrates for the aperture and the cavity back, respectively, $\varepsilon_{r1} = \varepsilon_{r2} = 3.55$ are the permittivities of the substrates, and $\lambda_0$ is the free-space wavelength. In this work, $H_1 = 25$ mm is chosen to avoid half-wavelength destructive interference ($D_0 < 0.5$) up to 4.88 GHz.

Since the cavity is electrically shallow below 1 GHz ($H_1 < 0.083 \lambda_0$), the reflective backing significantly affects the currents on the boom and LP aperture, resulting in deteriorated impedance match. To reduce the interaction between the cavity and the LP aperture and improve the match without increasing
A bowtie slot is etched on the cavity back underneath the boom as depicted in Fig. 5.35. The bowtie shape instead of a rectangle is adopted because it provides the larger slot area at the outer region where the interaction is stronger while the inner region has more reflective area to maintain the gain at high frequencies. The width of the slot can be controlled by the apex of each fan arm $B_2$. Fig. 5.36 shows VSWR of the antenna backed by the bowtie slot cavity for various $B_2$. This VSWR is much improved from that in Fig. 5.33 and can be further improved by increasing $B_2$ since the disturbance in the currents on the aperture is weakened. However, the slot reduces the gain and FBR as shown in Fig. 5.37, which is not surprising. Specifically, the LP antenna induces the current on the slot to be radiated to backward ($\theta = 180^\circ$) below 1 GHz. This unwanted back radiation is increased for the larger $B_2$. At the resonance frequency of the slot around 0.8 GHz, however, the slot backing reflects the back radiation of the LP aperture back to the forward direction so that the boresight gain and FBR are enhanced.
5.5.2 Terminated Slot Cavity with High FBR

To improve the gain and FBR, the radiation from the slot needs to be suppressed. Fig. 5.38 shows the electric field distribution in the slot at its resonance (0.76 GHz). The slot is essentially a magnetic dipole antenna and the strongest field is formed in the middle. Therefore, the slot is terminated by a lumped resistor, and an additional cavity with the same height is attached at the bottom as shown in Fig. 5.39. Fig. 5.40
shows VSWR of the antenna for various terminating resistance $R_{term}$. VSWR is much improved from Fig. 5.36 due to the suppressed radiation from the slot and can be further improved by increasing $R_{term}$. However,
the improved impedance match sacrifices the gain and efficiency as shown in Fig. 5.41. As a compromise between VSWR and gain, $R_{term} = 100\ \Omega$ is chosen. Fig. 5.42 shows the boresight gain and FBR for various lengths of the slot $L_{slot}$, which determines the resonant frequency of the slot. Though no major difference is seen in gain, FBR, and impedance match, $L_{slot} = 110\ \text{mm}$ is chosen to eliminate the gain drop at 0.8 GHz for $L_{slot} = 120\ \text{mm}$ and to achieve higher gain above 4 GHz.

Fig. 5.40. VSWR of the antenna in Fig. 5.39 for various terminating resistance $R_{term}$.

Fig. 5.41. Boresight gain and total efficiency of the antenna in Fig. 5.39 for various terminating resistance $R_{term}$. 

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Fig. 5.42. Boresight gain and FBR of the antenna in Fig. 5.39 for various $L_{\text{slot}}$.

Fig. 5.43. LP antenna backed by a bowtie slot terminated by a microstrip line and an absorber loaded additional cavity. (a) Overview. (b) Layout of the bowtie slot with microstrip termination.
Fig. 5.43 shows the configuration of the proposed cavity backed uni-directional LP antenna. The LP aperture is backed by the bowtie slot cavity that is terminated by a linearly tapered microstrip line, and the additional cavity is loaded with an absorber layer. The microstrip line uses one side of reflector as ground plane and is connected to the other side through a plated via. The lower end of the line is connected to a vertically mounted SMA connector on the bottom, where 50 Ω termination is used as shown. The microstrip line is linearly tapered from 3.33 mm to 0.4 mm, corresponding to 50 Ω and 100 Ω, respectively. By using the microstrip termination, the coupled power to the slot can be taken outside the cavity and recycled rather than dissipated as a heat. For instance, this coupled power can be used as a feeder into a rectifier subsystem designed to convert RF into DC. If improved impedance match is needed, an absorber layer may be laid in the additional cavity; however, some reduction in power handling is to be expected. Fig. 5.44 shows the VSWR of the proposed slot cavity backed LP antenna, slot cavity without the absorber in the additional cavity, and 25 mm single cavity backing filled with an absorber layer. The proposed antenna achieves VSWR below 2.5:1 from 0.59 GHz to 5 GHz. The impedance match of the proposed antenna is comparable to that of the absorber loaded single cavity backing above the turn-on frequency. Also, the electrical size of the antenna at the turn-on frequency is $0.49 \lambda_0 \times 0.49 \lambda_0 \times 0.11 \lambda_0$, much smaller than the horn antenna discussed in Section 5.1.

![Fig. 5.44. VSWR of the proposed antenna in Fig. 5.43, the proposed antenna without the absorber in the additional cavity, and the LP antenna backed by an absorber loaded cavity.](image)
Fig. 5.45. Boresight gain, peak gain, and FBR of the proposed antenna in Fig. 5.43, the proposed antenna without the absorber in the additional cavity, and the LP antenna backed by an absorber loaded single cavity.

Fig. 5.46. Simulated radiation patterns in (a) E- and (b) H-planes.

Fig. 5.45 shows the boresight gain, peak gain, and FBR of the three antennas. The proposed slot cavity backing has higher gain than single cavity over the most of bandwidth since the dissipated power by the absorber is smaller while FBR of the proposed cavity is as good as that of single cavity, demonstrating the good uni-directionality. Also, it is noticed that the peak gain is higher than the boresight gain above 3 GHz, indicating the direction of maximum radiation is not at the boresight. Fig. 5.46 shows the radiation pattern of the proposed uni-directional antenna. As shown, beamwidths are widened at 4 GHz and 5 GHz due to the destructive interference; however, the majority of the energy is still radiated toward the upper
hemisphere. Fig. 5.47 shows the total efficiency of the three antennas. For the single absorber cavity, the total efficiency ranges from 10% to less than 50% since the back radiation from the aperture is totally dissipated by the absorber. For the proposed slot loaded cavity, the total efficiency is improved and ranges from 30 to 80%. Additionally, the coupling losses to the microstrip termination with and without the additional absorber are included in Fig. 5.47. Without the absorber, -2.56 dB of input is coupled to the termination at 0.76 GHz (the resonance of the slot) while that is reduced to -3.82 dB for the proposed cavity due to the additional absorber. When input power to the antenna is high, this coupling loss can be recycled rather than dissipated to further improve the overall efficiency.

Fig. 5.47. Total efficiency and coupling loss to the terminated slot of the proposed antenna in Fig. 5.43, the proposed antenna without the absorber in the additional cavity, and the LP antenna backed by an absorber loaded cavity.

5. 6 Conclusion

A planar bi-directional LP antenna fed by a microstrip line with the novel wide-boom geometry is proposed and improved performance over its conventional embodiment in FD and ThD is demonstrated. Linked electro-thermal multiphysics simulation has shown that the proposed wide-boom considerably reduces abrupt temperature rise caused by the dipoles’ resonances. VSWR ≤ 2 and flat gain around 5 dBi
over a decade bandwidth were simulated and measured. To validate thermal modeling, an IR camera was successfully utilized to measure and then correlate temperature distribution on antenna under the CW input.

A wideband cavity backing to accomplish the uni-directionality is also designed. A slot under the boom helped with impedance match at low frequencies at the expense of increased back radiation. To enhance FBR, the slot is terminated by a microstrip line and an additional cavity loaded with absorber is attached. Microstrip line-based termination was developed to support versatile treatment of non-radiated power. The proposed uni-directional LP antenna achieved VSWR below 2.5:1, gain higher than 0 dBi, and total efficiency above 30%, and good uni-directional patterns over a very wide bandwidth.
CHAPTER 6

Conclusion

6.1 Summary

High power wideband antennas with omnidirectional, horizon-directional, bi-directional, and broadside-unidirectional radiation patterns for possible consideration in various electronic warfare (EW) systems are presented in this thesis. Proposed antennas while belonging to the different families of conventional antennas have several novelties specifically developed to make them capable of consistent operation over multiple physical domains. They are small and exhibit wide impedance and gain bandwidths, small dispersion, and high peak/average power handling capability.

To understand impedance properties of monopole-like antennas and help improve their performance, the equivalent circuit model of monopole, annular slot, and monocone are derived on physics-based observations related to their structure and electromagnetic fields. The proposed circuit models are shown to be valid over more than octave bandwidths for arbitrary dimensions of antennas. Empirical formulas that tie the circuit-level performance with the structural parameters are also derived and are of possible use for rapid initial and low-cost design and optimization. The developed circuit models are used throughout this thesis to minimize the impact of reflected pulse from a wideband monopole, reduce the electrical size of a monocone, and analyze the combination of monopole and annular slot modes.

New wideband millimeter wave omnidirectional and horizon-directional antennas are proposed for the receiver of an integrated towed radio decoy (TRD) system. To understand the impact of cylindrical decoy platform on antennas’ radiation performance, the careful computational study is conducted and extensive contour plots are used to better understand this multi-dimensional problem. It is shown that larger dimensions of cylinder in each of three principal planes result in the higher gain, smaller elevation angle squint, and narrower beamwidth. It is also seen that the surface currents on the cylinder can bring about significant edge diffraction which increases cross-polarization contamination when the cylinder is thin and long. Also, the surface current density can be used to predict the antenna isolation between multiple receiver
and transmitting antennas. The low-cost wideband connector antenna is manufactured to demonstrate the suitability of its performance on a decoy-like cylinder. This antenna is easily tuned to a quarter wavelength at the desired frequency and shows 38% bandwidth performance with consistent monopole-like patterns. When integrated with the cylinder, the connector antenna maintains its wide bandwidth and excellent agreement with the theory. To cover the entire bandwidth of interest for radar-guided missiles, aka 18 to 45 GHz, the magnetic loop current of the annular slot in the feeding coax is combined with the inherent electric monopole mode. The proposed combined annular slot-monopole antenna (CASMA) achieved VSWR ≤ 2, consistent gain around 5.5 dBi, and monopole-like patterns with electrical size of 0.12 $\lambda_0 \times 0.12 \lambda_0 \times 0.18 \lambda_0$. Next, a wideband offset paraboloid reflector backing for the CASMA is designed based on the phase center variation of the feeder to transform omnidirectional patterns into horizon-directed beam for the directional TRD receiver. Design parameters of the reflector are determined by an extensive parametric study, yielding good performance with VSWR ≤ 2.2:1, FBR > 20 dB, gain > 10 dBi, and consistent radiations patterns over the same bandwidth while the electrical size of reflector is kept at 0.60 $\lambda_0 \times 0.48 \lambda_0 \times 0.20 \lambda_0$. The small size, excellent performance, and low cost feature CASMA and reflector as a viable candidates for TRD receiver subsystems with omnidirectional or directional coverage. Moreover, performance of CASMA in time domain (TD) and power domain (PD) are assessed once it is scaled to lower frequency band. To improve power handling limitation due to the mismatch loss, the matching ring is designed using the equivalent circuit analysis and it is integrated with the CASMA. For PD considerations, a transition between a standard N-connector and the antenna aperture is designed. Excellent impedance match, moderate gain, and consistent patterns over a wide bandwidth along with good TD and PD performances render this antenna to be a suitable radiator for transient high power electromagnetics (HPEM) systems.

For transient HPEM systems with wider bandwidth, monocone antennas with and without a paraboloid reflector are demonstrated. The proposed antenna concept tackles critical issues in HPEM: reconfigurable pattern between omnidirectional and directional modes and beam scanning without moving the high power generator. For omnidirectional radiator/reflectors feed, a monocone is modified to a conical
monopole so that consistent monopole-like patterns over a wide bandwidth can be obtained with a reduced diameter. The reflector is designed to miniaturize its electric size while keeping an acceptable performance in FD and TD. To experimentally demonstrate the proposed concept, the reflector is fabricated and assembled with a mechanical scanning subsystem. The prototyped antenna has VSWR ≤ 2, gain > 10 dBi, FBR > 15 dB, and directional (in horizon) radiation patterns from 1.66 GHz to 20 GHz. The electrical size of the antenna at the turn-on frequency is only 0.383 λ₀ × 0.980 λ₀ × 0.324 λ₀, which is much smaller than typically used HPEM antennas. In addition, using an equivalent circuit models, a semi-helical loading is developed and electrical size of the monocone is miniaturized. The loaded monocone antenna achieves VSWR ≤ 2, consistent gain and monopole-like patterns from 1.23 (33.8% reduction) to 20 GHz.

Log-periodic (LP) antennas with bi- and uni-directional (broadside-directed) radiation characteristics are also researched in this thesis. A wide-boom modification of LP aperture is proposed to achieve good impedance and gain performances within a small size and maintained bi-directionality. The measured data of the LP antenna show good impedance match (VSWR ≤ 2), stable gain around 5 dBi, and bi-directional patterns over a decade bandwidth. The electrical size of the antenna is 0.36 λ₀ x 0.36 λ₀ x 0.02 λ₀. The multiphysics simulations and infrared camera are successfully utilized to characterize the antenna in electro-thermal domain numerically and experimentally, and the improvement in thermal behavior of the antenna due to the wide-boom geometry is demonstrated. The bi-directional pattern of LP antenna is then reconfigured to be uni-directional by adding a cavity backing with a high efficiency, small size, and light weight to possibly replace bulky and heavy ridged horn antennas. The proposed uni-directional antenna achieves VSWR ≤ 2.5, gain > 0 dBi, and total efficiency > 30% over 8:1 bandwidth.

In conclusion, the proposed antennas in this thesis can be adopted for a wide variety of EW systems requiring small size, wide bandwidth in FD, minimum distortion in TD, and high power handling capability in PD.

6.2 Contributions

The main contributions of this thesis include:
• Physically derived equivalent circuit model of monopole-like antennas including monopole, annular slot, and monocone is proposed and empirical equations for corresponding circuit parameters are provided. The proposed circuit models can be used for system level analysis as well as for the development of the similar kind of models for dipole and bi-conical antennas.

• Novel wideband monopole-based antennas are proposed for integrated towed decoy receiving system with omnidirectional and horizon-directional coverage. It is shown that the simple and low-cost millimeter-wave connector antenna achieves 38% bandwidth standalone or when integrated with a mockup decoy-like cylindrical platform. To achieve even wider bandwidth (18–45 GHz), the magnetic loop mode from the annular slot in the coax is combined with a monopole antenna in a single structure, named combined annular slot-monopole antenna (CASMA). Using the CASMA as a feeder, the wideband offset paraboloid reflector is designed and demonstrated experimentally. The additional design procedure is developed to ensure that the CASMA has good TD and PD performance for transient HPEM applications.

• Wideband monocone-based omnidirectional/horizon-directional antennas are proposed and considered for transient HPEM applications. A regular monocone is first modified to achieve consistent radiation over a wide bandwidth. It is shown that assigning 55% of antenna height as conical and the rest as cylindrical results in consistent monopole-like patterns over a wide bandwidth. Using the modified monocone as a feeder, a half-cut paraboloid reflector antenna is designed with mechanical scanning ability and pattern reconfigurability. The proposed antenna is characterized in TD and PD to assess its possible future use for transient high power EW applications. Finally, novel and simple miniaturization approach for omnidirectional monocone is proposed by loading semi-helical wires at the conical junction of the modified monocone.

• Novel wideband bi-directional and broadside-unidirectional LP antennas are proposed and considered for possible use in EW applications. The wide-boom geometry for planar bidirectional LP antenna is introduced to improve impedance match and gain. It is also demonstrated that this configuration also improves the thermal response of the antenna. Theoretical study of this antenna
performed in electro-thermal domain and the developed numerical models are validated experimentally. Additionally, the bi-directional LP aperture is configured for broadside-unidirectional operation by integrating with a semi-lossy cavity. This configuration allows not only for the antenna to be used for higher CW power applications but also to possibly recycle some of the lost non-radiated energy. The flush-mountable configuration of the proposed antenna was also shown to have light weight and high efficiency, thus indicating a way for future use at much lower frequencies in reasonable size packages.

6.3 Future Work

The work presented in this thesis can lead to new and interesting research:

- Dual-polarized receiver system for TRD can be implemented by using multiple CASMAs with and without the reflector along the cylinder's diameter. This may increase probability of detecting the enemy's radar signal. In this scenario, the isolation between antenna elements needs to be enhanced by adding corrugations, artificial impedance surfaces, and/or absorptive coating on the cylinder.
- Radome design and integration with CASMA is needed for environmental protection and to aid to the aerodynamics of the platform.
- Considering the performance of CASMA with and without the reflector in FD, TD, and PD, this antenna can be designed for transient HPEM applications with pattern reconfigurability and beam steering ability. Although the monocone-based reflector antenna has wider bandwidth, the CASMA-based reflector may be used for sub-hyperband (bandwidth ratio $(br) > 3$) HPEM systems with a smaller size and a better in-band impedance match. The design developed in this thesis is amenable to small modifications to make it capable of actual use in HPEM systems.
- The radiation pattern of the monocone fed reflector antenna has high side lobes due to the proximity of reflector to the feeder. If this performance is not acceptable for the system, the issue may be resolved by re-shaping the reflector and/or the feeder.
• Power handling capability of the monocone fed reflector antenna can be enhanced by using a customized connector and filling the high electric field region with gas/oil. In this regard, radome can be added to not only protect/conceal the antenna, but also enclose the filling medium.

• Integration of the semi-helical loading of the monocone with the reflector can further reduce the antenna size. Effect of self-resonance of the wire on the performance of reflector antenna in TD and PD should be considered.

• The flush-mountable uni-directional LP antennas can be considered for transient HPEM systems. This can be accomplished by investigating the effect of geometrical parameters of the antenna on TD and PD performance. Also, to achieve high peak power handling capability, a microstrip line needs to be replaced by a high power feed network.
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