Compact 3-bit Frequency Reconfigurable Monopole Antenna Realized with a Switchable Three-Line Section

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Abstract: This work proposes a compact 3-bit frequency-reconfigurable monopole antenna covering a broad reconfigurable range by inserting a switchable three-line section (STLS). The design starts with a conventional quarter-wavelength monopole line antenna, which is then replaced by a novel structure, the STLS. The STLS is composed of three parallel-connected lines with different lengths. Accordingly, three RF p–i–n diodes are introduced in the STLS to achieve binary reconfiguration. After all parameters of the antenna have been optimized, it will eventually output $2^N = 8$ (N is the number of switches) independent working states with different equivalent lengths and a reconfigurable working frequency. The number of states in a binary reconfigurable antenna is optimally large in relation to the number of switches used, which means that it can be extremely convenient for digital control of switching all the states and capable of decreasing the number of RF p–i–n diodes we used, thereby minimizing the manufacturing cost and loss of diodes. A prototype antenna is fabricated and tested, and the measurement results agree well with the simulation results, validating the good features, such as a large reconfigurable switchable frequency range from 0.95 GHz to 2.45 GHz with considerable working bandwidth varying from 40 MHz to 540 MHz for each state, simple structure, and a compact size of 70 × 40 mm$^2$, which can be appropriately used for a multi-radio wireless system and handheld devices. All the states have a similar monopole radiation pattern with a good maximum efficiency and an acceptable peak gain according to its compact size.

Keywords: frequency reconfigurable; microstrip monopole antenna; binary reconfiguration; RF p–i–n diode

1. Introduction

The antenna, as an indispensable part of the modern wireless communication system, has been continuously studied since it was invented many years ago. Reconfigurable antennas are becoming increasingly significant due to the multifrequency and multimode requirements of modern wireless communication and limited space for antenna occupancy [1–19]. These antennas possess several outstanding advantages, such as reconfigurability, size miniaturization, versatility, and a decrease in the number of antennas in a complex wireless system. The reconfiguration of an antenna can be roughly classified into frequency reconfiguration [1–17], polarization reconfiguration [18], and radiation pattern reconfiguration [19]. Frequency reconfiguration has been extensively investigated in the past few years.

Nowadays, a single radio device, especially handheld devices, may be required to handle many services over a wide frequency range and support various communications standards such as wireless local area networks (WLAN), Bluetooth, 3G, and 4G. A frequency-reconfigurable antenna (FRA) can typically cover a variety of bands for wireless communications, which is extremely applicable to handheld devices. Besides, unlike previous solutions using multi-frequency or ultra-wideband (UWB) antennas, which may introduce unwanted interference, FRAs offer noise rejection in bands that are not in use,
reducing the filter needs of the front-end circuits and resulting in an effectively larger bandwidth performance.

One of the most common FRAs possesses continuous frequency tuning agility by using varactor diodes [1–3]. A tunable U-slot antenna was proposed by inserting a varactor diode into the feed line of the antenna [1]. In [2], a frequency-tunable antenna based on quarter-mode substrate integrated waveguide (QMSIW) was presented by loading a varactor into the QMSIW but with a narrow working band from 2.28 GHz to 2.50 GHz. A pair of varactor diodes were mounted on the dipole arms to realize the frequency tuning agility [3]. A wide tuning range is difficult to construct with these frequency-tunable antennas.

The FRAs can also have a discretely switchable working frequency, and it may be realized by introducing switches, such as RF p–i–n diode switches [4–6,8–10,12–16] or microelectron-mechanical system (MEMS) switches [7,17]. In [4,5], frequency-reconfigurable microstrip patch antennas (MPAs) were presented with several different stubs connected to the main patch via switches. In [6], the frequency agility of the MPA was realized by switching shorting pins on and off on the patch, which can affect the input impedance. A conformal broadband patch antenna acquires its reconfigurability in frequency by connecting the main patch with extended edges via MEMS switches [7]. Frequency-reconfigurable microstrip slot antennas (MSAs) were proposed by reconfiguring the length of the slot [8] or of a loaded stub [9] by using switches, providing several different working states. A dipole antenna can also obtain frequency reconfigurability by reshaping its structure and dimensions using switches [10].

Microstrip monopole antennas (MMAs) are featured with low-profile characteristics and easy integration [20–22], and they have been adequately utilized for frequency reconfiguration [11–16]. A frequency-reconfigurable MMA with an ultrawideband mode was invented, and the reconfigurability was realized by using a slotted ground inserted with four p–i–n diodes, which has five different filtering states [12]. Six different operation modes are achieved using a triangular monopole antenna loaded with three stubs by adjusting the biasing condition of the three embedded diodes [13]. In [14], four p–i–n diodes are inserted on the antenna surface to vary the current distribution and adjust the resonant frequencies with different combinations of switches. The frequency reconfigurable MMA in [15] was presented by inserting three p–i–n diodes into the monopole patch to form several parasitic patches, providing four operating dual-band modes. A simple reconfigurable antenna with wide-band, dual-band, and single-band operating modes was designed by connecting two additional stubs to the triangular monopole through two p–i–n diodes [16].

However, most of these published FRAs can only switch within a narrow tuning range or provide a relatively limited number of working states while loading many switches and lumped biasing elements. Accordingly, binary reconfiguration can be applied to solve these problems by realizing $2^N$ switchable states ($N$ is the number of switches), which means that it can be extremely convenient for digital control of switching all the states. A few studies on multibit FRAs with binary reconfigurability have been recently published, achieving an optimally large number of states in relation to the number of switches used. In [17], a 3-bit frequency-switchable planar inverted-F antenna (PIFA) was investigated with the capability to switch with eight states by using three MEMS switches. However, it only covers a narrow frequency range from 1.52 GHz to 2.25 GHz.

In this work, a 3-bit frequency-reconfigurable antenna based on a conventional MMA is presented. Three RF p–i–n diode switches are allocated within a switchable three-line section (STLS), providing an optimally large number ($2^3 = 8$) of states of the MMA with different equivalent lengths and obtaining a binary frequency reconfigurability within a wide switchable frequency band.

The paper is organized as follows: Section 2 presents the antenna configuration and design process and theoretical analysis of binary reconfiguration. Section 3 shows the simulated results. Section 4 provides the measured results and discussion, and Section 5 concludes the study.
2. 3-bit Frequency Reconfigurable Antenna Design

(a) Antenna Configuration

A conventional simple quarter-wavelength MMA is selected as the basic structure for the antenna design because of its compact size and ease of integration. The geometry of the proposed antenna is shown in Figure 1, and the design is constructed on a Rogers RO4003C substrate with a relative permittivity ($\varepsilon_r$) of 3.38 and thickness (h) of 0.508 mm. Computer Simulation Technology (CST) software [23] is used to simulate the presented antenna and optimize the parameters. The size of the substrate is 70 $\times$ 40 mm$^2$.

![Figure 1. Geometry of the proposed antenna: (a) top and (b) back view.](image)

In contrast with the conventional MMA, an STLS, which contains three parallel-connected lines with different lengths, is inserted as part of the resonant branch. Each line of the STLS is split into two sections with an RF p–i–n diode switch connecting them. The lengths of each line can be given as $L_1 < L_2 < L_3$. There is a short connecting section between the STLS and the 50-ohm feed line, and there is an open-circuited section at the end. As each switch has two states, “On” and “Off”, the three switches provide $2^3 = 8$ working states in total for the FRA, and the antenna has a different effective length and working frequency in each state. The antenna is fed by a 50-ohm microstrip line with a width of $W_0$.

By turning on/off the switches within the STLS, we may obtain the states that can be named with binary numbers, and Table 1 lists all these states.

|       | 000 | 001 | 010 | 011 | 100 | 101 | 110 | 111 |
|-------|-----|-----|-----|-----|-----|-----|-----|-----|
| $D_1$ | OFF | ON  | OFF | ON  | OFF | ON  | OFF | ON  |
| $D_2$ | OFF | OFF | ON  | ON  | OFF | OFF | ON  | ON  |
| $D_3$ | OFF | OFF | OFF | OFF | OFF | OFF | OFF | ON  |

The RF p–i–n diodes from Skyworks with a P/N of SMP1331-079LF are used for switching [24]. Small gaps of 0.5 mm are reserved for placing the diodes and other lumped elements. A simple biasing circuit is designed for the three RF p–i–n diode switches.
Figure 2 illustrates the on and off state equivalent model of the p-i-n diode and schematic of the basing circuit.

![Biasing circuit](image)

**Figure 2.** On/off equivalent model of the SMP1331-079LF p-i-n diode and schematic of the basing circuit.

Figure 1a shows that three 100 pF capacitors, namely, $C_1$, $C_2$, and $C_3$, are introduced in the structure to realize DC-blocking at the input port and between every two lines of the STLS. Three biasing lines with a width of 0.3 mm are applied for 1 V DC supplies. Meanwhile, four 180 nH inductors, namely, $L_{in1}$, $L_{in2}$, $L_{in3}$, and $L_{in4}$, are used for RF choking. The schematic of the basing circuit in Figure 2 illustrates the manner by which the biasing circuit works. First, the DC voltage goes through the biasing line with an inductor at the end of it, which will prevent the surface current on the antenna from the DC current. Then, the DC voltage will pass through the antenna part to excite the diode. Accordingly, the DC current will flow to the ground through a via hole near the feedline.

### (b) 3-bit Reconfiguration Design

The 3-bit frequency reconfiguration is mainly introduced by the STLS. The connection status of STLS can be controlled by switching on or off the corresponding switches. In Figure 3, we have different equivalent lengths of the antenna related to the resonant frequency for all the eight states at the initial stage of the design, and they can be successively changed by switching to various states. The descriptions of the STLS for all the situations are discussed as follows:

![Equivalent structures](image)

**Figure 3.** Equivalent structures of the STLS for different states.

At state 000, all switches are turned off, and the equivalent length of the STLS is the shortest. Accordingly, the antenna works at the highest frequency. When each switch of the STLS is turned off, the corresponding line is split into two open stubs.

At states 001, 010, and 100, only one line of the STLS is connected into the resonating branch, and the working frequency of the antenna is mainly related to the length of the
line selected. The longer the line length, the lower the working frequency. We can obtain the following result due to \( L_1 < L_2 < L_3; f_{011} > f_{010} > f_{100} \).

The metal lines within the antenna may seem different from microstrip lines now that the grounding plane is removed. However, every small segment of the antenna part has its characteristic impedance, which is determined by the line width and the distance from the grounding plane of the feeding microstrip line. So, each line of the STLS may be regarded as a transmission line with high radiation loss and position-dependent characteristic impedance and, in addition, can be equivalent to a lossy transmission line with universal characteristic impedance and certain electrical length.

When two lines of the STLS are connected into the antenna, the rough equivalent circuit of the parallel connected lines can be equivalent to a single transmission line with a certain electrical length and characteristic impedance, affecting the working frequency for the three TLs are shown in Figure 5a,b by substituting the above equations into MATLAB for calculation. Variable \( \theta_e \) is between \( \theta_1 \) and \( \theta_2 \) with the change in frequency.

![Figure 4](image-url)  
**Figure 4.** Two parallel transmission lines can be equivalent to a new transmission line.

The \( Y \) matrices for TL1 and TL2 are as follows:

\[
[Y_1] = \begin{bmatrix}
-j \cot \theta_1 / Z_1 & j / Z_1 \sin \theta_1 \\
 j / Z_1 \sin \theta_1 & -j \cot \theta_1 / Z_1
\end{bmatrix}
\]  \hspace{1cm} (1)

\[
[Y_2] = \begin{bmatrix}
-j \cot \theta_2 / Z_2 & j / Z_2 \sin \theta_2 \\
 j / Z_2 \sin \theta_2 & -j \cot \theta_2 / Z_2
\end{bmatrix}
\]  \hspace{1cm} (2)

We can obtain the \( Y \) matrix of TLe as follows:

\[
[Y_e] = [Y_1] + [Y_2]
\]  \hspace{1cm} (3)

The \( ABCD \) matrix of TLe can be derived as follows:

\[
[ABCD]_{TLe} = \begin{bmatrix}
A_e & B_e \\
C_e & D_e
\end{bmatrix}
\]  \hspace{1cm} (4)

where:

\[
A_e = D_e = \frac{Z_1 \sin \theta_1 \cos \theta_2 + Z_2 \sin \theta_2 \cos \theta_1}{Z_1 \sin \theta_1 + Z_2 \sin \theta_2}
\]  \hspace{1cm} (5)

\[
B_e = \frac{jZ_1 Z_2 \sin \theta_1 \sin \theta_2}{Z_1 \sin \theta_1 + Z_2 \sin \theta_2}
\]  \hspace{1cm} (6)

Accordingly, we may compute \( \theta_e \) and \( Z_e \) with the following:

\[
\theta_e = \arccos A_e
\]  \hspace{1cm} (7)

\[
Z_e = -jB_e / \sin \theta_e
\]  \hspace{1cm} (8)

The corresponding curves of the electrical length and characteristic impedance versus frequency for the three TLs are shown in Figure 5a,b by substituting the above equations into MATLAB for calculation. Variable \( \theta_e \) is between \( \theta_1 \) and \( \theta_2 \) with the change in frequency,
and $Z_e$ is smaller than $Z_1$ and $Z_2$, which means that TLe has the distinct equivalent length and width from TL1 and TL2. The antenna frequency will be decided by $Z_e$, $\theta_e$, and the circuit parameters of other antenna parts. States 011, 101, and 110 will have broader bandwidths than states 001, 010, and 100, since $Z_e$ is smaller than $Z_1$ and $Z_2$. Moreover, for states 011, 101, 110, the resonating branch can be regarded as a stepped-impedance branch, and the working frequencies are pushed higher than those of states 001, 010, and 100.

**Figure 5**. Electrical length (a) and characteristic impedance (b) versus frequency for TL1, TL2, and TLe.

At state 111, all three lines of the STLS are connected. In this situation, the analysis of the equivalent electrical length and characteristic impedance of the STLS will be more complex, and electromagnetic (EM) software can provide accurate simulation results. The STLS will have the lowest equivalent characteristic impedance because all the lines are connected, introducing the broadest working bandwidth and the second highest working frequency.

(c) **Initial Design Method and Parametric Study of the Antenna**

Considering the design of an FRA covering a frequency range from the lowest frequency $f_L$ to the highest frequency $f_H$, we may know that $f_L$ and $f_H$ correspond to state 100 and state 000, respectively.

If neglecting the width differences between sections and neglecting the loading effects of the unconnected lines of the STLS, we may know the effective length of state 100 is $L_3 + L_4 + L_5 + L_6$, while the effective length of state 000 is about $L_6 + L_3/2$, according to Figure 3.

We may find the rough relationships that can be used for the initial selection of parameters:

$$L_3 + L_4 + L_5 + L_6 \approx \frac{c}{4f_L\sqrt{\varepsilon_{eff}}}$$  \hspace{1cm} (9)

$$L_6 + \frac{L_3}{2} \approx \frac{c}{4f_H\sqrt{\varepsilon_{eff}}}$$  \hspace{1cm} (10)

$$\varepsilon_{eff} \approx \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left(1 + 12 \left(\frac{h}{w}\right)\right)^{-0.5}$$  \hspace{1cm} (11)

where $c$ is the speed of light, $\varepsilon_{eff}$ is the effective dielectric constant, $h$ is the thickness of the substrate, and $w$ is the width of the microstrip lines.

Initially, we may also decide the length differences, $L_2 - L_1$ and $L_3 - L_2$, according to the differences between quarter-wavelengths:

$$L_3 - L_2 \approx \frac{c}{4f_H\sqrt{\varepsilon_{eff}}} - \frac{c}{4f_{010}\sqrt{\varepsilon_{eff}}}$$  \hspace{1cm} (12)
The ratio of the STLS length to the FRA length, namely, \( r_L \), also affects the working frequency range. If \( r_L \) is small, the structure reconfiguration of the STLS will not incur large range for frequency reconfiguration. In Figure 6a, when \( L_4 \), the length of the open-circuited section, is being increased and the STLS stays unchanged, \( r_L \) is becoming smaller and we can observe that the frequency of state 111 is decreasing very fast while the frequency of state 001 is decreasing very slowly, and the frequency ratio \( f_{111}/f_{001} \) is becoming smaller.

\[
L_2 - L_1 \approx \frac{c}{4f_{010}\sqrt{\varepsilon_{eff}}} - \frac{c}{4f_{001}\sqrt{\varepsilon_{eff}}}
\]  

(13)

The widths of all lines and sections also affect the working frequencies and bandwidths, since the antenna can be regarded as a stepped impedance resonating branch at all states. For example, when the width \( W_1 \) is changing from 1.5 mm to 2.55 mm, as shown in Figure 6b, the equivalent characteristic impedance of the STLS is becoming lower, and both frequencies of the 001 and 111 states are going higher. The working bandwidths are also increased when \( W_1 \) becomes larger as observed, and the impedance matching for the antenna is also improved.

A practical design of an FRA using STLS can be started by the initial selection of the lengths, widths, and considering the specified upper and lower working frequencies of the states. Optimization of the FRA by performing parametric simulation using commercial EM software, such as CST, can then be conducted for achieving desired the distribution of working frequencies and good impedance matching for each working band.

Following the upside methods, the example FRA is designed with the eight states, covering a wide frequency band from 0.95 GHz to 2.45 GHz. The FRA may be used for a reconfigurable communication system, since several important wireless communication standards, such as GSM900/1800, WiFi, Bluetooth, IEEE 802.11n, WCDMA and TD-LTE, etc., are included within the frequency range. The final optimized dimensions are determined as: \( L = 70 \text{ mm}, W = 40 \text{ mm}, L_0 = L_6 = 16.4 \text{ mm}, W_0 = 1.1 \text{ mm}, W_1 = 2.55 \text{ mm}, W_2 = 1.4 \text{ mm}, W_3 = 1.3 \text{ mm}, L_1 = 17.5 \text{ mm}, L_2 = 23.9 \text{ mm}, L_3 = 33.5 \text{ mm}, L_4 = 15.7 \text{ mm}, L_5 = 7.1 \text{ mm}, \) and \( L_6 = 5 \text{ mm} \).

### 3. Ultimate Layout Simulation on EM Software

In the practical design of the reconfigurable antenna, the RF p–i–n diode model is simplified to a resistance of 1.7 Ω and a capacitance of 0.18 pF at the on and off states for more efficient simulation, which can be implemented by the corresponding lumped elements on the CST software.

(a) Simulated current distribution

![Figure 6. Simulated characteristics of reflection coefficient at state 001 and 111 for different values of \( L_4 \) (a) and \( W_1 \) (b).](image-url)
Figure 7 illustrates the current distribution of the STLS for all of the states by CST simulation, where it can be confirmed that the STLS works as described in Figure 4. Figure 7a demonstrates that state 000 has the lowest current distribution on the STLS because all the switches are turned off, and the current will not flow through the diodes. Figure 7c shows the STLS at state 010, which has the densely distributed current in the left section because switch $D_2$ is active and will cause the current to mainly flow into line two. Additionally, when switched to state 111 in Figure 7h, the uniform distribution of the surface current occurs on the STLS because three switches are turned on.

![Simulated surface current distribution of the STLS at different states](image)

**Figure 7.** Simulated surface current distribution of the STLS at different states. Panels (a–h) correspond to states 000–111.

**b)** Reflection coefficient simulation

Figure 8 illustrates the simulated reflection coefficient results at different states of the proposed antenna. The eight states are totally independent and have good impedance matching performance that covers a very large switchable frequency range from 0.97 GHz to 2.40 GHz (for each state, only the first resonant mode is accounted). The return loss, $S_{11}$, is better than 10 dB within the operating frequency band. At the lowest working frequency (0.97 GHz), the proposed antenna reached a very compact dimension of $0.226 \lambda_0 \times 0.129 \lambda_0$, where $\lambda_0$ is the wavelength at this frequency.

![Simulated reflection coefficient](image)

**Figure 8.** Simulated reflection coefficient at different states of the proposed 3-bit frequency reconfigurable MMA.
In addition, certain unexpected harmonics arise for several states, such as states 000 and 001 already marked in Figure 6, which are mainly caused by the split open stubs of the STLS when switches are turned off, and it can be practically solved by using some narrowband filters or utilizing some harmonic suppression techniques, such as bringing a defected ground structure orthogonal to the feedline [25] and adding an open stub at a suitable position away from the feed point [26].

4. Measured Results and Discussion

(a) Measured reflection coefficient results

The measured results of the reflection coefficient and a photograph of the antenna prototype are provided in Figure 9a,b. The $S_{11}$ results are measured with an Agilent N5245A vector network analyzer, and the antenna can switch between 0.95 GHz and 2.45 GHz. Good impedance matching is obtained for each state, and the simulated and measured results of the working frequency and operating bandwidth at each state are all shown in Table 2.

Figure 9. Measured reflection coefficient at different states (a) and photograph (b) of the fabricated 3-bit frequency reconfigurable MMA.

Table 2. Simulated and measured resonance frequency and operating bandwidth for all the eight narrowband states.

| State | Simulated Resonance Frequency (GHz) | Measured Resonance Frequency (GHz) | Simulated Operating Bandwidth (MHz) | Measured Operating Bandwidth (MHz) |
|-------|------------------------------------|------------------------------------|-------------------------------------|-----------------------------------|
| 000   | 2.23                               | 2.27                               | 360                                 | 400                               |
| 001   | 1.28                               | 1.23                               | 70                                  | 150                               |
| 010   | 1.04                               | 1.00                               | 35                                  | 40                                |
| 011   | 1.44                               | 1.41                               | 100                                 | 130                               |
| 100   | 0.98                               | 0.97                               | 25                                  | 45                                |
| 101   | 1.60                               | 1.52                               | 190                                 | 190                               |
| 110   | 1.70                               | 1.62                               | 370                                 | 290                               |
| 111   | 1.88                               | 1.79                               | 470                                 | 540                               |

Good agreement between the simulation and the measurement results can be observed, except for the small frequency shift of approximately 53 MHz on average for each state. The measured bandwidths for most of these states become larger than the simulated ones as a result of the unavoidable deviation of antenna machining accuracy and the introduction of
DC voltages. Additionally, the p–i–n diode model is simplified in the simulation, which will also cause discrepancies between the measured and the simulated results. Actually, there is an inductor series-connected to the resistance or the capacitance that we purposefully ignored due to its detrimental influence on simulation efficiency. Figure 9a shows some perturbations on the reflection coefficient curves, which are caused by the parasitic effects of the biasing lumped elements.

(b) Far-field result discussion

In the far-field situation, the proposed antenna must maintain the similar bidirectional radiation characteristics in the E-plane (xz-plane) and nearly omnidirectional pattern in the H-plane (yz-plane) for all the eight states, which accord with the typical printed MMA far-field patterns, as shown in Figure 10. This work has been tested in an anechoic chamber. Figure 11 shows the simulated and measured far-field results at states 001, 101, and 111, where various distortions have been observed on the measured radiation patterns due to the DC supplies for the three switches and possibly the parasitic effects brought by the lumped elements.

Figure 10. Simulated 3D radiation pattern of states 001 (a), 101 (b), and 111 (c) at 1.28 GHz, 1.60 GHz, and 1.88 GHz, respectively.

Figure 11. Simulated and measured radiation patterns at their corresponding resonant frequencies of states 001 (1.28 GHz and 1.23 GHz), 101 (1.60 GHz and 1.52 GHz), and 111 (1.88 GHz and 1.79 GHz): (a) E-plane; (b) H-plane.
The simulated far-field results still agree well with the measured output. Cross-polarizations have been effectively suppressed. The simulated gain for each state varies from $-1.25$ dBi to $2.3$ dBi, and the antenna efficiency is between $56\%$ and $98\\%$, as shown in Figure 12. State 000 reaches the maximum gain and efficiency because all diodes were switched off, and state 100 has the minimum gain due to its low efficiency. Moreover, the measured peak gain and efficiency are $1.5$ dBi and $94\%$, respectively, because the p–i–n diodes and the lumped elements have introduced some losses. The losses may be effectively reduced by replacing the p–i–n diodes with MEMS switches.

![Graph showing simulated and measured gain and efficiency at the resonant frequencies of different states.](image)

**Figure 12.** CST simulated and measured gain and efficiency at the resonant frequencies of different states.

(c) Comparison and improvement methods

The proposed 3-bit MMA has a more compact size [5,6,8,13,16], larger average BW [5,6,8], and larger number of switchable working states [5,6,8,13,15,16] compared with those published works due to the binary frequency reconfigurability (Table 3). This work, as a binary reconfigurable antenna, has a larger switchable frequency range and larger working bandwidth than the 3-bit binary reconfigurable antenna recently presented [17].

| Ref. | Type | Size ($\lambda^2$) | No. of Switches | No. of States | Binary Re-configurable | Switching Range | Aver. BW | Peak Gain (dBi) | Efficiency |
|------|------|-------------------|----------------|---------------|------------------------|-----------------|----------|----------------|------------|
| [5]  | MPA  | $0.39 \times 0.39$ | 6              | 36            | x                      | 37\%            | 1.7\%    | 4.3            | 73\%       |
| [6]  | MPA  | N/A               | 16             | 8             | x                      | 37\%            | ~3\%     | 7.8            | 94\%       |
| [8]  | MSA  | $0.58 \times 0.53$ | 5              | 6             | x                      | 73\%            | 8.4\%    | >1.9           | N/A        |
| [13] | MMA  | $0.54 \times 0.27$ | 3              | 6             | x                      | N/A             | N/A      | 3.31           | >80\%      |
| [15] | MMA  | $0.35 \times 0.21$ | 3              | 4             | x                      | 105\%           | 8.2\%    | 3.85           | 89\%       |
| [16] | MMA  | $0.46 \times 0.28$ | 2              | 3             | x                      | N/A             | N/A      | 4.1            | 92\%       |
| [17] | PIFA | $0.65 \times 0.65$ | 3              | 8             | √                      | 39\%            | 2.7\%    | 4.5            | 95\%       |
| This work | MMA | $0.38 \times 0.22$ | 3              | 8             | √                      | 88\%            | 13.5\%   | 1.5            | 94\%       |

* $\lambda$ is the wavelength of the center frequency. Switching range and average bandwidth (Aver.BW) are not given in reference [16,19] because they have wideband modes.
The peak gain of the reconfigurable antenna is slightly lower than those of other solutions, and this is a common weakness of microstrip monopole antennas [11,20–22], which have a bidirectional radiation pattern in the E-plane and an omnidirectional radiation pattern in the H-plane and are more suitable for wireless communication devices, especially handheld devices. The low efficiency, particularly in state 100, is mainly caused by p–i–n switch diodes, which also decreases the gain of the antenna. The efficiency may be improved by choosing switches with lower loss or by optimizing the loading positions of the STLS and the p–i–n diodes for reducing ohmic losses, enhancing the gain accordingly. The antenna gain could be effectively improved if a reflector is added under the antenna, restraining the EM radiation to half space. Although the peak gain is not high, the reconfigurable MMA obtains more miniaturized dimensions than reconfigurable patch antennas, following the performance tradeoff of antennas.

5. Conclusions

In this study, a compact 3-bit frequency reconfigurable MMA realized with an STLS has been proposed. As a binary reconfigurable antenna, the number of states is optimally large in relation to the number of switches used, allowing for digital control of the operating frequency band and decreasing the number of p–i–n diodes used. The measured results agree well with the simulated ones, and a good impedance matching characteristic and large bandwidth are achieved on each narrowband state, covering a very wide frequency range from 0.95 GHz to 2.45 GHz (88% switching range) with considerable bandwidth varying from 40 MHz to 540 MHz. Furthermore, this antenna has similar far-field radiation characteristics to the conventional MMA with a maximum efficiency of 94% and an acceptable peak gain of 1.5 dBi, and it also features compact dimensions. Above all, the proposed 3-bit frequency reconfigurable MMA is extraordinarily suitable for applications in handheld devices, which are required to handle many services over a wide frequency range and support various communications standards such as wireless local area networks (WLAN), Bluetooth, 3G, and 4G.

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