Multiport analysis and design of segmented antennas

Darko Kajfez¹ | Roger Hasse² | Lucia Scialacqua³

¹Department of Electrical Engineering, University of Mississippi, Oxford, Mississippi, USA
²Lockheed Martin Space Company, Littleton, Colorado, USA
³Microwave Vision Group (MVG), Rome, Italy

Abstract
A design procedure is described which enables simultaneous control of the input-impedance match, directivity, and bandwidth of segmented antennas. Their impedance properties can be accurately determined by a straightforward multiport analysis, which makes it possible to optimise the values of individual reactances with the use of closed-form objective functions. Two practical omnidirectional segmented monopole antennas have been designed, built, and tested, to verify the proposed design procedure. One is a single-tuned design with the narrow bandwidth of 1% and realised gain of 6.7 dBi, and another is the double-tuned design with 42% relative bandwidth and realised gain of 6.2 dBi.

1 | INTRODUCTION

The segmented antennas discussed here are a class of antennas consisting of short conductor segments (no longer than one-quarter wavelength), interconnected with lumped inductances or capacitances. They can be easily fabricated by printed-circuit techniques in form of loops or monopoles [1–4].

By changing the values of individual reactances of a segmented antenna, the designer is influencing not only the input impedance, but also the radiation pattern of the antenna system. We want to emphasise that, in contrast to the segmented antennas described elsewhere [5–7], we allow each reactance to get a distinct value, thereby giving the designer considerably more degrees of freedom. The values of individual reactances are found by optimisation. The optimisation to be described here is performed instantly on any laptop computer, allowing the designer to investigate various scenarios in a short amount of time [8].

A solid wire monopole antenna over an infinite ground plane produces a perfect omnidirectional radiation pattern. The lowest resonance occurs when the monopole is one-quarter wavelength long, while the corresponding directivity in the horizontal direction is 5.1 dBi. The second resonance occurs when the monopole is one-half wavelength long, and its directivity is two times larger, namely 8.1 dBi. Unfortunately, the solid half-wavelength monopole antenna has an extremely high input impedance, which is difficult to match to 50 Ω. The advantage of properly designed segmented antennas is that they are matched to the impedance of a feeding cable, so that an external matching circuit is not needed.

Design procedures described in [1–4] provided narrow-band operations, with relative bandwidths from 3% to 6%. The novel design procedure described here achieves considerably wider and/or considerably narrower 10 dB bandwidths, as documented in Section 5.

Section 2 defines the equivalent circuit for a segmented monopole antenna. Sections 3 and 4 summarise the proposed design procedure that enables a trade-off between the bandwidth, directivity, and matching of segmented antennas.

Experimental verification of the proposed design procedure is documented in Section 5. Two practical half-wavelength segmented monopoles are presented: a narrowband antenna, and a wideband antenna. They can be classified as single-tuned and double-tuned designs [9].

2 | CIRCUIT ANALYSIS OF THE SEGMENTED ANTENNA

As shown in Figure 1, the half-wavelength monopole antenna investigated here consists of six segments of equal length 10.5 mm, separated by gaps of 1 mm.

The design objective is to operate at the centre frequency of 2.44 GHz, so that its total length of 69 mm is slightly longer...
than one-half of the free space wavelength \( \lambda_0 \). The segments are 3.6 mm wide and assumed to be made of a perfect electric conductor (PEC). It is important to realise that any segmented antenna constitutes a multiport, which is driven at one or more ports, while the remaining ports are loaded by pure reactances. Therefore, antenna behaviour may be described by an impedance matrix \( Z_a \). When the dimensions of the antenna are specified, the impedance matrix can be obtained by an electromagnetic simulation software (EMSS). Our impedance matrix was generated by the software FEKO [10].

The right-hand illustration of Figure 1 represents the general equivalent circuit of a segmented antenna. The monopole type circuit will have only one voltage source \( V_{g1} \) and the other voltage sources \( V_{g2} \) to \( V_{g6} \) will be zero. The reactances \( jX_i \) (i = 1–6) will be either capacitances or inductances. They form a diagonal matrix:

\[
Z_{ex} = \text{diag}(0, jX_2, ..., jX_6).
\]

For practical reasons, we chose the reactance \( X_1 = 0 \), so that the centre pin of an RF connector can be directly soldered to the first segment of the antenna. The network description of the system is

\[
|I\rangle = (Z_a + Z_{ex})^{-1} |V_g\rangle,
\]

where the vector \( |I\rangle \) contains the currents \( I_i \) on the individual ports, and the vector \( |V_g\rangle \) contains the voltage sources at all ports. The solution of the matrix Equation (2) will provide the values of the currents on all the ports, and also the input impedance as well as the total input power \( P_{in} \) delivered to the antenna.

A full description of the antenna properties requires also the knowledge of its radiation pattern. More specifically, one is interested in the knowledge of the antenna directivity \( D \). Since each segment of the antenna is shorter than one-quarter of wavelength, the current distribution along the segments cannot display any standing waves, but it can only gradually vary between one port and the next one. Assuming a linear variation, the effective segment current that contributes to radiation can be approximated by:

\[
I_{si} = (I_i + I_{i+1})/2,
\]

where \( i = 1–6 \), and the current at the top of the sixth segment is set to be zero.

Now we have enough information to express the directivity \( D \) of the monopole antenna by using basic antenna theory [11]:

\[
D(\theta) = \frac{30}{P_{in}} \left( \frac{\pi}{\lambda_0} \right)^2 \sum_{i=1}^{6} |I_i d_i \sin \theta e^{ijkz \cos k} |^2.
\]

In the above expression \( \lambda_0 \) is the free-space wavelength, \( d_i \) is the length of the \( i \)-th segment, \( z_i \) is the vertical position of the midpoint of the \( i \)-th segment, and \( k \) is the propagation constant of the free space. The angle \( \theta \) is the polar angle of spherical coordinates. This explicit equation for evaluating the radiation pattern can be executed instantly on any laptop computer. Although the present communication discusses only the segmented monopole antenna, the same procedure to evaluate approximate directivity can be applied to all segmented wire antennas (dipoles, loops, polygons etc.) that can be described by the equivalent circuit shown in Figure 1. An explicit expression for the directivity pattern of a half-loop segmented antenna can be found in [3].

3 | SINGLE-TUNED DESIGN PROCEDURE

The bandwidth, return loss, and directivity of the segmented antenna all depend directly on the choice of reactances connected to the individual ports. Maximising the value of the antenna Q factor will achieve the minimum bandwidth. Control of the impedance bandwidth thus becomes an optimisation problem, where an optimal combination of five reactances has to be found so that a desired bandwidth is achieved, while maintaining the best trade-off for the return loss and the directivity.

The bandwidth of a single-tuned antenna is inversely proportional to its Q factor. The value of the antenna Q factor can be estimated from the behaviour of the input impedance \( Z_{in} \) as a function of frequency [12]:

\[
Q = \frac{\int_0^f \frac{dZ_{in}}{df}}{2 \text{Re}(Z_{in})}.
\]

This equation should be applied in the vicinity of the segmented antenna resonant frequency \( f_0 \).

The objective function for optimisation consists of three parts:

\[
U = w_u U_u + w_D U_D + w_Q U_Q.
\]
The input match is controlled by the function $U_m$ and the corresponding weight $\omega_m$, the directivity is taken care by the function $U_D$ and weight $\omega_D$, while the value of $Q$ is controlled by the function $U_Q$ and the weight $\omega_Q$. The initial weights can all be set to unity and the expected value of each partial objective function have also been made close to unity, when it is within the acceptable range of values. Next, the target values $\rho_T, D_T$, and $Q_T$ should be selected, for the expected values of the input reflection coefficient magnitude $\rho$, directivity $D$, and the antenna $Q$ factor. Then, the first partial objective function is defined as follows:

$$\text{if } \rho \leq \rho_T \rightarrow U_m = 0, \quad \text{else } U_m = \left(\frac{\rho}{\rho_T} - 1\right)^2. \quad (7)$$

The reasoning for the above definition is the following. As long as the reflection coefficient is smaller than its target value, the matching objective function $U_m$ is set to zero. This provides an opportunity for the other two partial objective functions to find their optimal values. When the reflection coefficient becomes twice as large as the target value, $U_m$ should become unity. Afterwards, it grows towards much higher values. The directivity-related partial objective function $U_D$ is defined as follows:

$$\text{if } D \geq D_T \rightarrow U_D = 0, \quad \text{else } U_D = \left(10\left(1 - \frac{D}{D_T}\right)\right)^2. \quad (8)$$

Therefore, if the directivity falls to 90% of the targeted value, $U_D$ becomes unity. If the directivity decreases even further, the value of $U_D$ grows fast towards 100. The third partial objective function $U_Q$ assures the low value of $Q$:

$$\text{if } Q \leq Q_T \rightarrow U_Q = 0, \quad \text{else } U_Q = \left(\frac{Q}{Q_T} - 1\right)^2. \quad (9)$$

Thus, if $Q$ is twice as large as the target value, $U_Q$ becomes unity, whereas beyond that value it grows rapidly.

For the first optimised design, the weight $\omega_Q$ was set to zero in order to obtain a high-directivity narrowband operation at the frequency 2440 MHz. To check the accuracy of the approximate evaluation of directivity, the EMSS model is comprised of PEC strips and of ideal, lossless lumped inductances and capacitances. The values of individual reactances have been found using the minimisation procedure “minminax” from the Matlab® Optimization Toolbox [13]. The acceptable range of reactance values was set to be between $-500$ and $+500 \Omega$. The resulting radiation pattern as a function of the angle $\theta$ consists of a main lobe at $\theta = 90^\circ$, and a side lobe about 8 dB lower, as shown in Figure 2.

![Figure 2](image.png)  
**Figure 2** Comparing the approximate with the accurate method of directivity evaluation. EMSS, electromagnetic simulation software.

The proposed design procedure can be summarised as follows:

1. **Step 1**: Generate the antenna impedance matrix at required frequency (for the narrow-band design), or several frequencies (for the wide-band design), with the use of an EMSS software.
2. **Step 2**: Determine values for all needed reactances by an available optimisation procedure.
3. **Step 3**: Compute the approximate dimensions of printed-circuit capacitances and inductances by simple static formulae.
4. **Step 4**: If the preliminary results look promising, enter these data into the EMSS software as the starting point of the full-blown optimisation of detailed antenna dimensions and materials.

Step 1 requires the use of the EMSS software. Steps 2 and 3 are executed instantly, so they can be repeated many times, while experimenting with the most suitable combination the input impedance, bandwidth and directivity. When also antenna dimensions need to be modified, repeating the first three steps does not require any significant loss of time. The execution of the Step 4 requires the use of the EMSS routine, so that each optimisation would require considerably longer time. This step is optional and even may not be needed if the first three steps provide enough accuracy for the problem at hand. For the single-tuned antenna design, Step 4 was executed only once, with the idealised perfect conductors, and by using static approximations for the lumped capacitances and inductances. This was done to verify the accuracy of the approximate closed-form Equation (4) for the directivity.

**4 | DOUBLE-TUNED DESIGN PROCEDURE**

The design procedure for the double-tuned antenna follows the same steps, but the particular objective functions for the control of bandwidth are selected in the following way. For a
double-tuned response, the reflection coefficient contour should exhibit a small loop, situated close to the centre of the Smith chart. This desired behaviour is enforced by starting from the values of the impedance matrices evaluated by EMSS at five frequencies covering the intended bandwidth. It is required that the five points on the Smith chart lie inside a prescribed radius (e.g. $\rho_T = 0.316$ that corresponds to a return loss of 10 dB). To assure the creation of a loop, it is also required that the first and the last frequency points are close to each other, thus forming a loop. Therefore, the objective function for matching is defined as

$$U_m = \sum_i U_{mi} + U_{co} \quad (10)$$

All five of the $U_{mi}$ functions are defined in analogy with Equation (7), while the additional function $U_{co}$ is intended to assure a crossover of the lowest and the highest frequency points:

$$\text{if } \left| \frac{\Gamma_1 - \Gamma_5}{\rho_T} \right| < 1 \rightarrow U_{co} = 0 \text{ else } U_{co} = 3 \left( 30 \left| \frac{\Gamma_1 - \Gamma_5}{\rho_T} \right| \right)^2 . \quad (11)$$

Instead of being based on the antenna Q, the third part of the overall objective function is now defined to be a ratio of the largest reflection coefficient to the smallest directivity:

$$U_r = \frac{P_{\text{max}}}{D_{\text{min}}} \quad (12)$$

The partial objective function related to directivity is a sum of five values, one for each of the five frequencies:

$$\text{if } D_i < D_T \rightarrow U_{D_i} = 0 \text{ else } U_{D_i} = (2(D_i - D_T))^2 . \quad (13)$$

The double-tuned example to be described next, has been obtained with the following settings: $\omega_m = 1$, $\omega_r = 1$, $\omega_D = 1$, $D_T = 4$, $\rho_T = 0.25$. The five selected frequencies were: 2.14, 2.34, 2.54, 2.74 and 2.84 GHz. In the Step 4, the full optimisation was not performed. Instead, the dimensions obtained in the first three steps were directly substituted in EMSS for verification.

The magnitude of the return loss for a double-tuned response agrees very closely with the results of the fast design procedure, as displayed in Table 1. The table also shows that the values of directivity in the horizontal direction, obtained by the fast procedure (Steps 2 and 3) compared with those obtained by the EMSS are also in a satisfactory agreement: at the five design frequencies, the differences are all smaller than 1 dB.

### Table 1 Comparing the optimised and simulated results

| $f$ (GHz) | 2.14 | 2.34 | 2.54 | 2.74 | 2.84 |
|-----------|------|------|------|------|------|
| Return loss (dB) | | | | | |
| Optimised | 13.99 | 10.26 | 11.05 | 15.61 | |
| Simulated | 13.91 | 10.30 | 10.98 | 15.46 | |
| Directivity (dB) | | | | | |
| Optimised | 5.81 | 6.22 | 6.31 | 5.72 | |
| Simulated | 6.07 | 7.05 | 7.15 | 6.55 | |

5 | EXPERIMENTAL RESULTS

For fabrication of the prototype antennas, a Rogers 5880 substrate of relative dielectric constant $\varepsilon_r = 2.20$, and of thickness 0.787 mm was selected. The capacitances were created by overlapping the ends of neighbouring segments on the opposite sides of the substrate. The electrostatic formula for parallel-plate capacitors was used to determine the area of the overlapping regions. The required inductances were created by small circular loops interconnecting the neighbouring segments. The magnetostatic formula for the inductance of the circular loop [14] was used to determine the dimensions of the printed circuit loops. The resulting printed circuit prototypes of the single-tuned and the double-tuned segmented antenna are shown in Figure 3.

In the left photograph one can see two circular-loop inductances. The apparently missing segments are printed on the opposite side of the substrate. On the right-hand photograph the two relatively large circular plates, together with two of equal size on the opposite side, were printed to form the relatively large capacitance that was needed at port 2.

The measured amplitude of the reflection coefficient for the single-tuned antenna, displayed in Figure 4, shows that the narrowband antenna has the resonant frequency equal to 2415 MHz, about 1% lower from the design frequency of 2440 MHz. The frequency shift is believed to be caused by the use of the approximate static formulae for capacitances and inductances.

The detailed view of both printed-circuit antennas is shown in Figure 5. Each antenna is printed on both sides of the substrate. The overlapping parts for capacitive loading are visible as darker areas within dotted borders. Also, notice that the wideband antenna has only one inductance added at port 5, and it is split into two parallel circular loops. The two loops cancel each other’s radiation, such that the resultant pattern of the wideband antenna is omnidirectional, with very low cross-polarised radiation (i.e. 30 dB below peak amplitude).

The EMSS model of the single-tuned antenna built with ideal PEC conductors shows that its radiation Q factor would be as high as $Q = 536$. This high value might be perhaps approached with superconductors. But, from the measured value of the input reflection coefficient, the antenna Q factor was estimated to be only $53 \pm 8$ [15]. It is interesting that the conductor and dielectric losses have made the antenna Q 10 times smaller than a computed value that would have been achieved with lossless materials.

The measured and simulated input reflection coefficients for the double-tuned prototype are shown on the Smith chart in Figure 6. As can be seen, the contours lie within the dotted
circle indicating the 10 dB return loss. To align measured and simulated contours with each other, the actual measured contour had to be rotated by an amount corresponding to 11 mm in free space, which is the effective length of the SMA coaxial connector.

The magnitudes of the input reflection coefficients are shown in Figure 7. It can be seen that the measured value is not as well matched as the idealised (lossless) design value, but the 10 dB bandwidths agree with each other within 5%.

The prototypes were measured in a spherical near-field chamber at MVG laboratories [16]. During the measurement, antenna was mounted on a circular ground plane of 1 m diameter.

As the prototypes were designed to operate above an infinite ground plane, the measured radiation patterns from one-m ground plane were post-processed by a technique described in [17], in order to reconstruct the appearance of each antenna radiation pattern on an infinite ground plane. In practice, diffractive contributions from the edges of the finite ground plane are removed reproducing the radiation on an infinite ground plane. Then by using the inverse source technique [18], that reconstructs the equivalent currents on a 3D geometry representing the antenna only, all the remaining contributions (not radiated by the antenna) are spatial filtered.

**Figure 3** Narrowband (left) and wideband (right) prototypes of segmented antennas

**Figure 4** Measured (solid line) and simulated (dashed line) reflection coefficient magnitudes of the narrowband antenna

**Figure 5** Overlapping details of the narrowband (left) and the wideband (right) prototype antennas
The radiation pattern of any monopole antenna should be omnidirectional. As can be noticed in Figure 8, the measured gain of the single-tuned antenna in the horizontal plane is gradually varying between 5 and 7.5 dBi. This variation is believed to have been caused by radiation from relatively large inductance loops at ports 3 and 5.

In all figures that follow, the solid line denotes the measured value, and the dashed one denotes the simulated value. The vertical radiation pattern for the single-tuned antenna is shown in Figure 9. As shown in Figure 5, the antenna has two sizeable inductance loops at ports 3 and 5. When the antenna current passes through these loops, some undesired radiation is created. Each loop forms an elementary magnetic moment, oriented perpendicularly to the surface of the loop. We believe that this cross-polarised radiation is the reason for filling the theoretical null in the z direction in Figure 9. If the cross-polarised radiation is considered detrimental for a specific application, the inductance values should be selected smaller, and each inductance should be split into two equal inductances, like those of the double tuned antenna (see the right-hand side of Figure 5). For a distant observer, the two closely spaced antiparallel magnetic moments will cancel each other.

The measured omnidirectional coverage for the double-tuned antenna departs from an ideal circle by less than 1 dB, so it will not be shown here. The vertical radiation patterns for the double tuned antenna shown in Figures 10–12 are taken at
TABLE 2  Measured values of realised gain and efficiency

|               | Gain dBi | Efficiency | f MHz |
|---------------|----------|------------|-------|
| Single-tuned  | 6.69     | 0.85       | 2415  |
| Double-tuned  | 4.65     | 0.81       | 1890  |
| Double-tuned  | 6.19     | 0.92       | 2440  |
| Double-tuned  | 4.74     | 0.77       | 2950  |

TABLE 3  Comparing the bandwidths and gains of various segmented omnidirectional antennas

|               | 10 dB BW | Max. gain |
|---------------|----------|-----------|
| Single-tuned  | 1%       | 6.7 dBi   |
| Double-tuned  | 42%      | 6.2 dBi   |
| Reference [5] | 66%      | 4.2 dBi   |
| Reference [6] | 9%       | 2.7 dBi   |
| Reference [7] | 9%       | 1.8 dBi   |

measured directivity of this segmented antenna is 7.4 dBi. This value comes quite close to the maximum theoretical value of 8.1 dBi, mentioned in the Introduction.

We are not aware of other segmented antennas that operate on the same principle as described in this communication. Reference [5] is a segmented dipole antenna, while references [6, 7] are segmented circular loop antennas. Table 3 compares the percent bandwidths, and maximum gains of various segmented antennas. It can be seen that our wideband prototype has 2 dB larger measured maximum gain, but 24% narrower relative bandwidth than the wideband antenna in reference [5]. Antennas from references [6, 7] show considerably lower gains.

6  | CONCLUSIONS

Partitioning the conductors of a printed circuit antenna into a number of short segments enables controlling the bandwidth, from a very narrow to a relatively wide value. A fast design procedure, based on optimisation, allows one to control the bandwidth, while simultaneously maintaining the target values for the return loss and the directivity of the antenna. Validity of the proposed design procedure has been confirmed with experimental data. The same design procedure can be applied to other segmented antennas configured as loops, half-loops, dipoles, monopoles, or arbitrary polygons.

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ORCID

Darko Kajfez  https://orcid.org/0000-0002-6668-2513
REFERENCES

1. Hasse, R., Demir, V., Hunsecker, W., et al.: Design and analysis of partitioned square loop antenna. Appl. Comput. Electromagn. Soc. J. 23(1), 53–61 (2008)
2. Kajfez, D., Elsherbeni, A.Z., Demir, V., et al.: Omnidirectional square loop segmented antenna. IEEE Antenn. Wireless Propag. Lett. 15, 846–849 (2016)
3. Nayeri, P., Elsherbeni, A.Z., Hasse, R., et al.: Half-loop segmented antenna with omnidirectional hemispherical coverage for wireless communications. Appl. Comput. Electromagn. Soc. J. 1(3), 88–91 (2016)
4. Kajfez, D.: Segmented wire antennas. Elektrotehniški Vestn. 83(4), 177–182 (2016). Available online. http://ev.fe.uni-lj.si/online.php
5. Sun, H, Ding, C.Y., Guo, Y.J., et al.: A wideband dipole antenna based on a non-uniformly segmented structure. In: Proceedings of the 11th European Conference on Antennas and Propagation, pp. 3572–3574. Ljubljana, Slovenia (2017)
6. Ma, Z.L.: Inductively loaded segmented loop antenna by using multiple radiators. IEEE Antenn. Wireless Propag. Lett. 65(16), 109–112 (2011)
7. Quing, X., Chen, Z.N.: Horizontally polarized omnidirectional segmented loop antenna. In: Proceedings of the 6th European Conference on Antennas and Propagation, pp. 2904–2907 (2011)
8. Nayeri, P., Kajfez, D., Elsherbeni, A.Z., et al.: Design of a segmented half-loop antenna. In: Proceedings of the IEEE Antennas and Propagation Society International Symposium, Memphis, July 2014
9. Lopez, A.R.: Double-tuned impedance matching. IEEE Antenn. Propag. Mag. 54(2), 502–508 (2012)
10. FEKO Suite 6.2: EM Software & Systems-SA (Pty) Ltd., 32 Technopark, Stellenbosch, 7600, South Africa (2012)
11. Collin, R.E.: Radiation from simple sources. In: Antenna Theory, Part I, pp. 21. McGraw-Hill, New York (1969)
12. Kajfez, D., Wheeless, W.P.: Invariant definitions of the unloaded Q factor. IEEE Trans. Microw. Theor. Tech. 34(3), 840–841 (1986)
13. Matlab®: Mathworks, Inc.: Natick, MA. Available online. http://www.mathworks.com
14. Ramo, S., Whinnery, J.R., Van Duzer, T.: Fields and Waves in Communication Electronics, pp. 311. John Wiley & Sons, New York (1965)
15. Kajfez, D.: Q Factor Measurements Using Matlab®, pp. 69–78. Artech House, Boston (2011)
16. MVG Inc.: 2105 Barrett Park Dr., Suite 104, Kennesaw GA, 30144. Available online. http://www.microwavevision.com
17. Foged, L.J., Misek, F., Benevenga, B., et al.: Infinite ground plane antenna characterisation from limited ground plane measurements. In: Proceedings of the 2010 IEEE Antennas and Propagation Society International Symposium. Toronto, Canada (2010)
18. Araque, J.L., Vecchi, G.: Improved accuracy source reconstruction on arbitrary 3-D surfaces. IEEE Antenn. Wireless Propag. Lett. 8, 1046–1049 (2009)

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