

Chapter 2

Parasitic Antenna Arrays: The Antenna Perspective

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Abstract This chapter provides the parasitic antenna dimension as an enabler of MIMO transmission, showing how antenna design, RF engineering, and MIMO processing can be used in parallel in order to achieve such a challenging goal. The main mechanism of creating beam patterns in the analogue domain with a single active and multiple parasitic antennas is described, as well as the associated methods to control the generated beams/waves. The dependence between beam-shaping and MIMO transmission for transmit diversity is presented, paving the way for MIMO transmission via parasitic antenna arrays. The relationship of parasitic array MIMO to conventional MIMO is also introduced, revealing fundamental similarities despite the differing approaches.

Keywords Antenna • Parasitic antenna array • Angle of arrival • Mutual coupling

2.1 The Smart Antenna Divide

Smart antenna systems have emerged in the early 1970s [1], as a means to calculate the angle of arrival (AoA) of incoming radio signals for military applications. In these applications, a radio frequency receiver has to determine the direction wherefrom a hostile radio transmission is initiated. The basic concept is taking advantage of the propagation delay of such transmissions, assuming that the signal propagates through the air as a plane wave. Under this assumption, if the receiver

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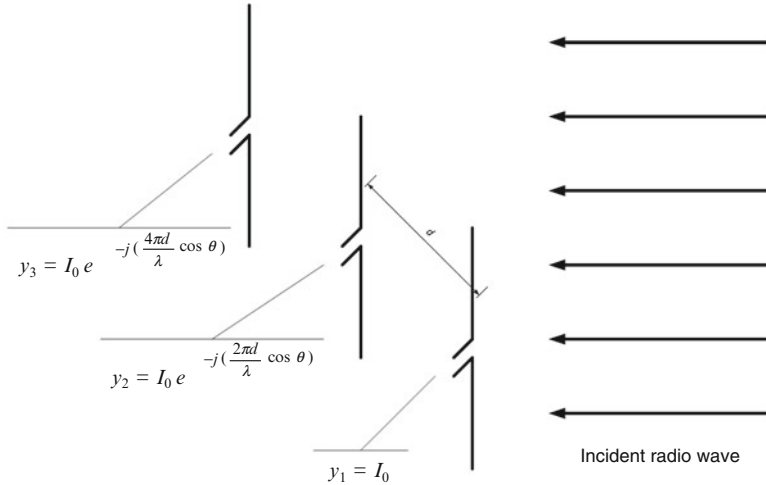


Fig. 2.1 A linear antenna array and an incident plane wave

is equipped with multiple antennas that sample the same signal at different points in space, then the AoA of the incident wave can be derived by calculating the time that the incident wave hits each antenna element. An example is shown in Fig. 2.1, where a linear antenna array is considered. However, the same concept is used for planar arrays or even three-dimensional arrays.

Such receivers that were able to independently sample incident waves from different antenna elements were called “smart antennas.” This definition, which has largely affected the way that we design smart antennas up to this day, has also dictated the way that we conceive smart antenna systems, based on the following assumption: *Antennas elements are dumb. They are just passive devices that transform an electrical signal that lies on our circuits, to an electromagnetic signal that propagates in the air.* Consequently, since antennas themselves cannot be smart, one needs an intelligent system to process the signals coming from the antenna elements and combine them in such a way, so as to achieve a predefined goal: AoA estimation, interference cancellation, beam-forming and beam-steering, etc.

Therefore, it is well established that in order to be able to construct a smart antenna, one needs to have multiple antenna elements, be able to acquire their signals in an independent fashion, and process them in the digital part of the circuit. This led to a system design paradigm, which was in essence the integration of multiple super-heterodyne receivers, one for each independent antenna element. This design paradigm, which has not been altered for decades, consists of the radiating element part, the RF/IF part, and the digital signal processor (DSP), as shown in Fig. 2.2. Due to the large capabilities and the advantages of digital signal processing, the analogue stages of the design have minimal processing capabilities,

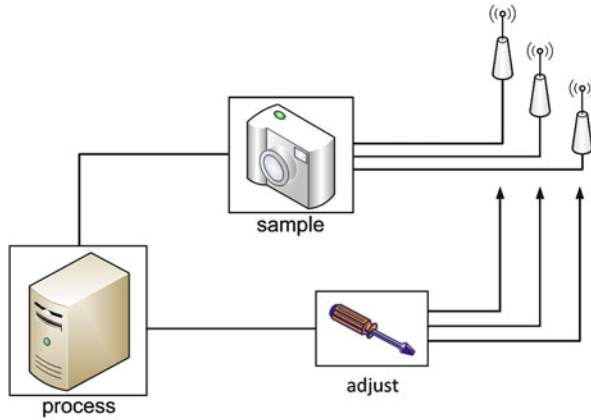


Fig. 2.2 Conceptual diagram of a smart antenna system

restricted to some analogue filtering and switching between transmitter and receiver paths. Therefore, in smart antenna designs, the main functionality of the analogue stages of the receiver is to simply translate the RF signals into comprehensive baseband signals for processing in the digital part of the circuit.

This simple architecture, which has dominated the design of smart antenna systems, has led to the decorrelation of antenna research and signal processing research. This is what we call the “smart antenna divide.” On the antenna side, the goal in smart antenna systems design is to produce antenna arrays capable of providing independent signals to the digital systems for processing. The digital signal processing side has no concern of the antenna design itself, but rather models the antennas as sampling points in the spatial domain.

This analytical approach to the design of smart antennas has proved to be quite efficient for decades, and has produced remarkable smart antenna systems. Examples of such systems are direction of arrival estimation and high accuracy beam-forming systems, which take advantage of the developments in digital signal processing technology. MUSIC [2, 3] and ESPRIT [4, 5] algorithms, blind source separation algorithms, and SDMA algorithms have increased the performance of cellular systems. However, with the proliferation of cellular telephony, and the urge for smaller and more power efficient handheld devices, the analytical design approach has reached its limits. It seems that, in order to advance the performance of smart antenna systems, we need to reconsider their design paradigm, bridging the gap between the antenna design world and the digital signal processing world.

In this chapter we initiate this effort by reviewing some basic concepts, trying to solve the main problem when bringing different worlds together: language and definitions.

2.2 Basic Antenna Concepts

Following different research paths with diverse design goals, the antenna design and the signal processing communities have developed their own terminology, making it quite hard to communicate ideas and concepts from one world to the other. For example, antenna engineers are more interested in mutual-coupling between antenna elements, while signal processing experts focus on signal cross-correlation. Antennas should have high radiation efficiency and maximize received signal power, while the signal processing algorithms utilize complex representations of signals in order to maximize spectral efficiency. Consequently, terms like bandwidth, gain, patterns, and efficiency have different meaning and value for the antenna engineer and the signal processing expert. Here we review some of the basic concepts involved in smart antenna design, in order to provide a common ground for terminology and clarify any misunderstanding of commonly used terms.

2.2.1 Antennas as Circuit Devices

Antennas are devices that transform electrical signals that propagate within conductors, to electromagnetic waves that propagate in the air. As such, they can be considered as part of a circuit device, and therefore they can be characterized by metrics such as their impedance and radiation resistance. The antenna impedance is a complex number that describes the relationship between the voltage and the current at the antenna port:

$$Z_{\text{in}} = \frac{V_{\text{in}}}{I_{\text{in}}} \quad (2.1)$$

where V_{in} and I_{in} are complex numbers representing the amplitude and phase of the voltage and current on the antenna port, respectively. The antenna impedance is a useful metric, since it can be used to directly import the antenna as a device into any electronic circuit. Consider for example the simple circuit shown in Fig. 2.3, where a voltage source is connected to an antenna. Using the circuit equivalent, we may determine the desired antenna port impedance value for which the transfer

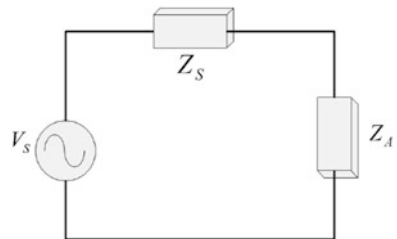


Fig. 2.3 Circuit equivalent of antenna and single source

of energy from the source to the antenna is maximized. In the example shown in Fig. 2.3, circuit theory dictates that when conjugate matching is applied, i.e.,

$$Z_A = Z_S^* \quad (2.2)$$

the source will transfer maximum power to the antenna. For any other antenna impedance value, some of the source power will not be absorbed by the antenna, but will be reflected back to the source. In antenna terminology this effect is called impedance mismatch, and antenna designers use the following metric to identify whether the antenna is properly matched to the circuit:

$$R_L \text{ (dB)} = -20 \log_{10} |\Gamma| \quad (2.3)$$

where

$$\Gamma = \frac{Z_L - Z_S}{Z_L + Z_S} \quad (2.4)$$

R_L is called return loss and it is the logarithmic equivalent of the reflection coefficient of the antenna. The meaning of return loss is the following: if $RL=0$ then all energy is reflected back from the antenna. If $R_L = -20$ dB, then only 1% of the incident energy is reflected back. In antenna design we consider that we have good matching properties when $R_L < -10$ dB.

Note that the value of the antenna impedance is directly related to the antenna geometry and to the environment in which the antenna is placed. Therefore, it is a property of the device and its environment, not of the signal that enters the device. In other words, if an antenna is poorly matched, no linear transformation of the signal can improve the antenna matching. Furthermore, since the input impedance (and therefore the matching efficiency) of the antenna depends on the environment in which the antenna is used, the antenna environment should be taken into account during the design process. Apple's i-Phone4 designers could elaborate on the consequences of failing to do exactly that.

Finally, the antenna impedance is also frequency dependent. This means that the antenna does not maintain the same impedance value for all input frequencies. Therefore, the antenna will produce a different return loss value for each frequency component of the signal. These values are often used to determine the bandwidth of the antenna. For example, in Fig. 2.4, the return loss of a 6 cm dipole antenna is shown against input frequencies. The bandwidth of the antenna is defined as the spectrum of frequencies for which the return loss is less than -10 dB. In the example of Fig. 2.4, this corresponds to a bandwidth of 200 MHz, or 10.52% of the resonant frequency.

The return loss and consequently the reflection coefficient Γ of such graphs are nominally calculated for source impedances equal to 50Ω . This is done mainly for historical reasons. However, if the actual source impedance of the targeted design is different, the antenna designer should take this into account, targeting at the maximization of the transferred power, according to Eq. (2.2).

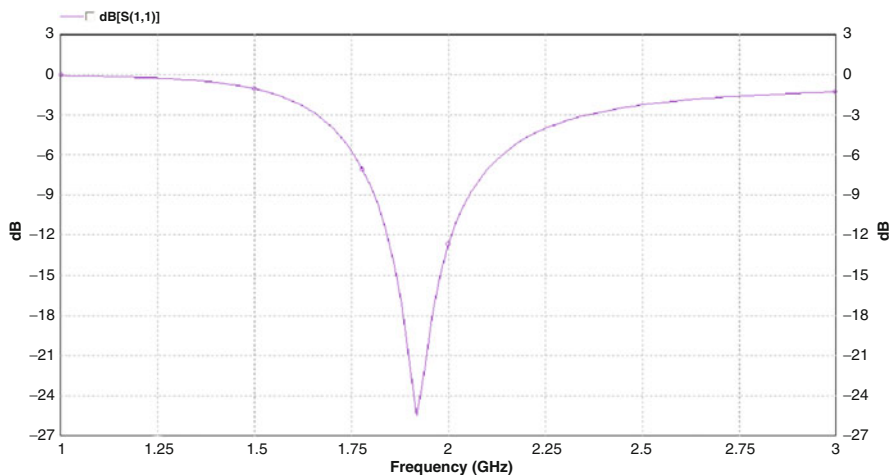
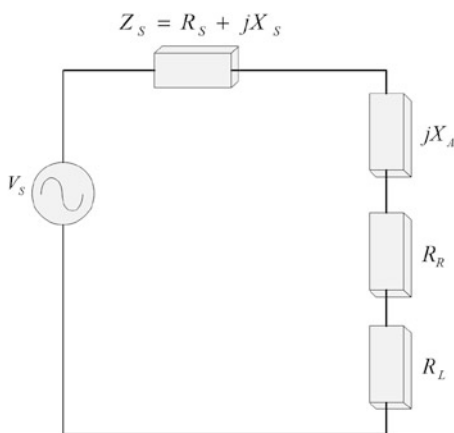


Fig. 2.4 Return loss of a printed 6 cm dipole antenna

Fig. 2.5 Circuit equivalent of antenna and signal source, including radiation resistance and resistive losses



Not all energy absorbed by the radiating element is transformed into electromagnetic waves radiated to the air. Some of the energy that enters the antenna is simply absorbed by the antenna material and dissipated in the form of heat. An example of this effect is shown in Fig. 2.5. In the circuit equivalent of the antenna these losses are modeled as resistive losses. On the other hand, the energy radiated into the air is modeled as a “radiation resistance.” The sum of these two resistance values constitutes the real part of the antenna impedance.

The goal of the antenna designer is to maximize the radiated power and minimize ohmic losses. Therefore, the radiation resistance and the resistive losses of the antenna play a significant role in antenna design. Since they represent signal power

that is either radiated or lost, they are used to determine another critical antenna metric which is called antenna efficiency:

$$\eta = \frac{P_r}{P_r + P_{\text{loss}}} \quad (2.5)$$

The antenna efficiency describes the percentage of power that the antenna actually radiates. It must not be confused with matching efficiency. For example if an antenna has a return loss of -10 dB (meaning that 90% of the power sent to the antenna is used) and a radiation efficiency of 50%, this means that only $90\% \cdot 50\% = 45\%$ of the power reaching the antenna is transformed into electromagnetic waves propagating into the air. Equivalently, we should expect that the same antenna when used at the receiver side would transform only 45% of the incident power into useful signals that the receiver could process. It is therefore evident that the aforementioned metrics play a significant role in antenna and communications systems design, since they dictate the actual power that is transferred from a wireless transmitter to a wireless receiver.

2.2.2 *Antennas as Electromagnetic Radiators*

The key function of antennas is to radiate electromagnetic waves to the three-dimensional space surrounding them. Therefore, the key metrics that characterize antennas as radiators are related to the form of the electromagnetic waves created and to the spatial distribution of the radiated power. Electromagnetic waves are formed by the intervention of time varying electric and magnetic fields, a phenomenon that is described by the well-known Maxwell's equations. In terrestrial communications, we tend to classify these waves according to the orientation of the electric field with respect to the earth plane. The orientation of the electric field may be parallel to ground, it could be vertical to ground, or it could be rotating around an axis. We would therefore classify an electromagnetic wave as having a horizontal polarization, vertical polarization, or circular polarization. Antennas are therefore classified according to the polarization of the electromagnetic waves they produce. An antenna that produces horizontally polarized waves is called a horizontally polarized antenna. An antenna that produces a vertically polarized wave is called a vertically polarized antenna, etc. We care about antenna polarization mainly because an antenna design which is optimized for a single polarization will only be efficient in receiving waves of the same polarization. For example, vertically polarized antennas will receive the vertical component of an arbitrary polarized wave, while horizontally polarized antennas will receive the horizontal component, etc. This is a significant property of antennas, and it has been used in the past to increase the capacity of wireless communications, or to decrease the interference of co-located communication systems.

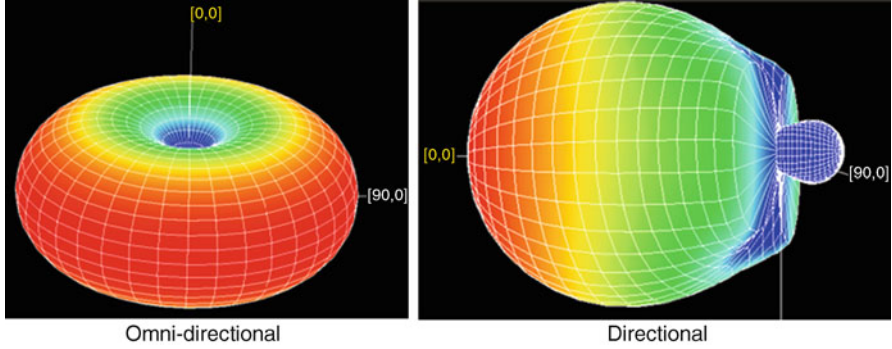


Fig. 2.6 Omni-directional and directional radiation patterns

A second set of metrics of electromagnetic radiators is related to the way that the radiated electromagnetic power is distributed in the three-dimensional space. Accordingly, we may classify antennas into three main categories: isotropic antennas, omni-directional antennas, and directional antennas. Isotropic antennas are those which emit the same amount of electromagnetic power toward all directions in space. Omni-directional antennas are those which emit the same amount of electromagnetic power toward all directions on a single two-dimensional plane in space. Directional antennas concentrate the emitted electromagnetic power toward a limited number of directions in space. We may identify which category an antenna lies within, once we measure the amount of power that this antenna sends toward all directions in space.

The isotropic antenna is mainly a very useful reference antenna, rather than a real antenna. It is also the type of antenna traditionally considered when developing signal processing algorithms for smart antenna systems. Nevertheless, antennas that have realistic matching and radiation efficiency values will not radiate in the same fashion toward all directions in space. The isotropic antenna is mainly used as a point of reference for describing the spatial distribution of an antenna's radiated power in spherical coordinates, using a metric called directivity. We therefore define directivity of an antenna toward a single direction (ϑ, φ) as the ratio of radiation intensity toward that direction, to the radiation intensity averaged over all directions, i.e.,

$$D(\vartheta, \varphi) = \frac{U(\vartheta, \varphi)}{U_0} = 4\pi \frac{U(\vartheta, \varphi)}{W_{rad}} \quad (2.6)$$

where W_{rad} is the radiation power. The three-dimensional chart that illustrates the directivity values of an antenna toward all directions in space is called the “radiation pattern” of the antenna. Examples of radiation patterns of the three aforementioned antenna categories are shown in Fig. 2.6.

Taking into account the polarization of the field emitted from the antenna, we may also define different radiation patterns per polarization. For example, a linear

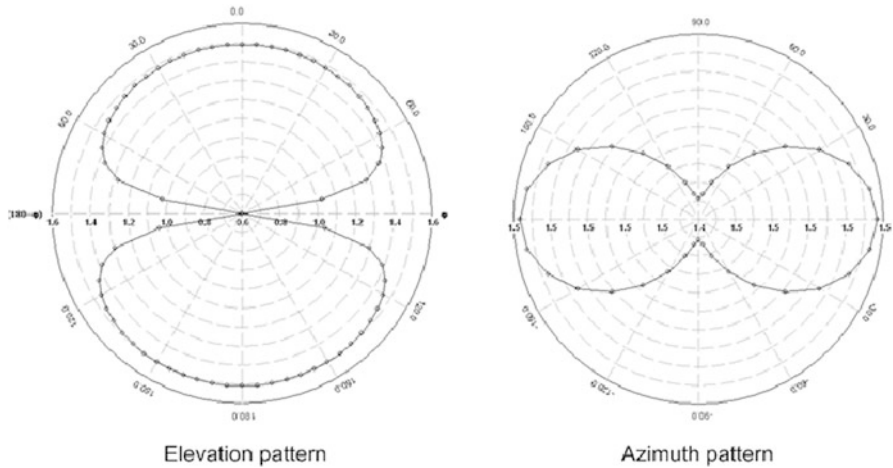


Fig. 2.7 Co- and cross-polarization patterns of a linear dipole antenna

dipole antenna placed vertically with respect to the ground plane will mainly radiate vertically polarized waves into the air. This is the principal polarization of the antenna and the pattern that describes the spatial distribution of the power of this wave is the co-polarization pattern of the antenna. However, not all emitted power has the same polarization. In the dipole example, some of the emitted waves do have horizontal polarization. The pattern that shows the spatial distribution of the power of these waves is called the cross-polarization pattern. These patterns are often superimposed in the same chart, as shown in Fig. 2.7.

As mentioned above, directivity is used to describe the spatial distribution of an antenna's radiated power. However, as described in the previous paragraph, not all the power that is fed in the antenna is transformed into electromagnetic waves. In order to have a picture of the three-dimensional distribution of emitted power, taking into account antenna losses, the metric of antenna power gain is used, where,

$$G(\vartheta, \varphi) = \eta \cdot D(\vartheta, \varphi) \quad (2.7)$$

One of the fundamental characteristics of antennas is that antenna gain is directly related to antenna size. Small antennas cannot achieve high gains. One needs to have antennas of significant size in order to achieve high gains. This is described in the following equation:

$$G_{max} = \eta \frac{4\pi A}{\lambda^2} \quad (2.8)$$

where $G_{max} = G(\vartheta, \varphi)|_{max}$, λ is the wavelength of the radiated signal, and A is directly related to the area that the antenna structure occupies. Note that the size of the antenna directly depends on the wavelength of the signal that the antenna

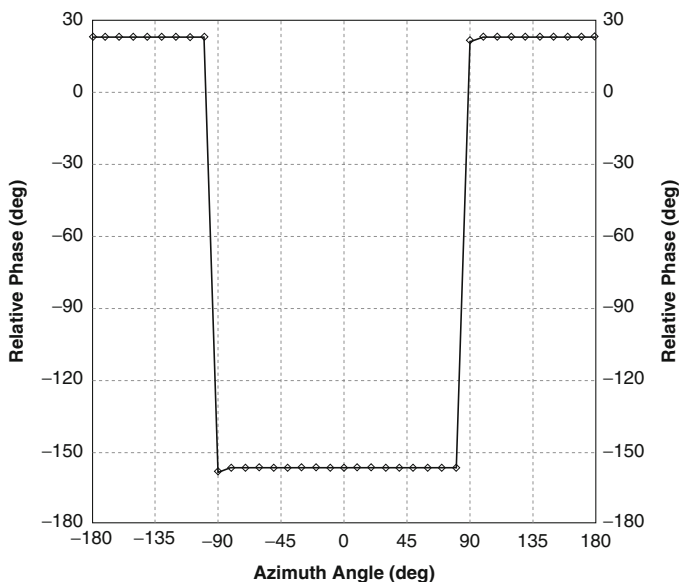


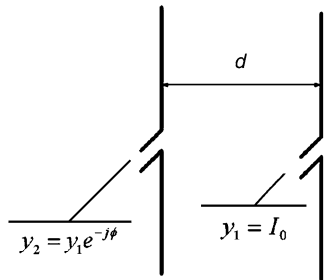
Fig. 2.8 Phase pattern of a rectangular patch antenna

radiates. If we have no information on the frequency of the signal, we cannot classify the antenna as small or large. For example, a metallic structure that has a length of 15 cm, is a small antenna when we feed it with a signal of 100 MHz, and it is classified as a large antenna when we feed it with a signal of 10 GHz. There are various rules of thumb for considering an antenna to be electrically small. In this book we are primarily concerned with antenna structures that have maximum length in the order of half a wavelength.

It is noted that the metrics of directivity and gain are directly related to the transmitted signal power, and are widely used in antenna design. Another metric that specifies the transmitted signal form, and has been largely overlooked for many years, is the phase lag of the transmitted field components behind its value at a specified reference point. This metric is quite useful for the beam-space analysis of transmitted signals followed here, considering that the reference point is the antenna feed point. It can be used together with the power characteristics of the transmitted signal to fully characterize the transmitted waveforms toward all directions in space. The three-dimensional graph that illustrates these relative phase lags toward all directions in space is called the “phase pattern” of the antenna, an example of which is shown in Fig. 2.8.

There are quite a few other metrics that describe an antenna as a radiator (front-to-back ratio, side-lobe level, field regions, phase center, etc.). However, in order to follow the beam-space analysis of this book, the aforementioned metrics, namely polarization, directivity, gain, and radiation pattern, are adequate.

Fig. 2.9 An adaptive two-element dipole array



2.2.3 Antenna Arrays and the Curse of Mutual Coupling

Note that all the aforementioned metrics are used for all antenna types. From the antenna perspective, the polarization, directivity, gain, and radiation pattern have the same meaning regardless of the antenna design used. The antenna designer has therefore the freedom to use any antenna structure to achieve the design goals related to the aforementioned metrics. For example, a high gain antenna can be designed as a single high-gain radiating element such as a helix or dish antenna; or it could be structured by forming a large array of closely spaced dipoles or patch antennas. Although the radiation characteristics of the two diverse design approaches could be engineered to coincide, in the latter case the antenna structure would have to have more than one input ports. This key difference adds to the adaptability as well as to the complexity of the array design. On one hand, an antenna array would be able to control its radiation characteristics on the fly, by altering the amplitude and phase characteristics of the signals fed to the different antenna ports. For example, the two-element dipole array shown in Fig. 2.9 could concentrate the transmitted signal power either to the left or to the right of the page, by simply changing the phase difference between the two feeds.

On the other hand, the design of antenna arrays poses additional constraints to the antenna designer: power fed to one of the array elements is induced to the other array elements. This could have two destructive effects on the antenna performance. Part of the induced power could leak back to the transceiver through the antenna ports or it could be scattered within the antenna structure, circulating among different antenna elements. In both cases, the ratio of radiated power over the total power fed to the antenna decreases, severely affecting antenna matching and radiation efficiency. This effect is called mutual coupling, and it is modeled using microwave network theory. In the example of the two element array of Fig. 2.9, if we view the input ports as ports of a microwave network, we can model mutual coupling using an impedance matrix:

$$v_i = Zi_i \quad (2.9)$$

where i_i are the currents fed to the input ports, and v_i are the total voltages due to antenna feeds and mutual coupling. The non-diagonal terms Z_{ij} represent mutual

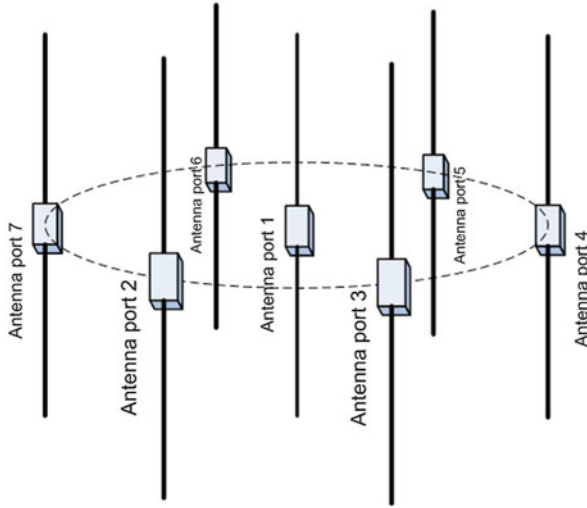


Fig. 2.10 A typical circular array

coupling, showing the voltage induced in port i , when port j is fed with current equal to $i_j = 1$ A. Note that the self-impedance of port i , Z_{ii} , is not the same as the input impedance of the same element, Z_0 , when used as a stand-alone element. For an N element array, (2.9) becomes

$$\mathbf{v}_A = \mathbf{Z}_A \mathbf{i}_A \quad (2.10)$$

where $\mathbf{v}_A$ is a vector of voltages across the element ports and $\mathbf{i}_A$ is a vector of the element port input currents.

As is the case with the input impedance, the impedance matrix describes the inherent structure of the antenna array and does not depend on the signals that are fed to the different antenna elements. Take also into account that in most antenna array implementations, $Z_{ij} = Z_{ji}$, meaning that the impedance matrix is usually symmetric. Further symmetries in the impedance matrix form usually depict symmetries in the antenna array structure. Take for example a circular array, as the one shown in Fig. 2.10. The form of the impedance matrix in that case would then be

$$\mathbf{Z} = \begin{bmatrix} Z_{00} & Z_{01} & Z_{01} & Z_{01} & Z_{01} & Z_{01} & Z_{01} \\ Z_{01} & Z_{00} & Z_{12} & Z_{13} & Z_{13} & Z_{15} & Z_{16} \\ Z_{01} & Z_{21} & Z_{00} & Z_{12} & Z_{13} & Z_{13} & Z_{26} \\ Z_{01} & Z_{31} & Z_{32} & Z_{00} & Z_{12} & Z_{13} & Z_{13} \\ Z_{01} & Z_{41} & Z_{42} & Z_{43} & Z_{00} & Z_{12} & Z_{13} \\ Z_{01} & Z_{13} & Z_{52} & Z_{53} & Z_{54} & Z_{00} & Z_{12} \\ Z_{01} & Z_{12} & Z_{13} & Z_{63} & Z_{64} & Z_{65} & Z_{00} \end{bmatrix}, \quad (2.11)$$

where in addition to the array's symmetry, identical dipoles are assumed, implying $Z_{ii} = Z_{11}$. As already noted, mutual coupling has been traditionally considered as a curse in antenna array design, mainly due to the fact that coupled power is fed back the circuit as a reflected wave, decreasing the efficiency of the design. This effect becomes more severe as we bring antenna element closer together. Many researchers in the field of antenna array design have tried to overcome mutual coupling problems, either by trying to isolate antenna elements on the array by intuitive antenna design, or by introducing complex feeding networks, negating the induced voltages at the antenna ports. However, it has been shown in [6] that for very small inter-element distances (in the order of $\lambda/10$), none of these approaches come without having to pay a price on antenna efficiency. Therefore, typical antenna arrays have inter-element distances in the order of $\lambda/2$ or higher, making the integration of such arrays into mobile terminals quite improbable.

2.3 Antennas in Communication Systems

The main function of antennas is to radiate waves carrying information into the air, and transmit them over large distances. The basic antenna concepts described in the previous paragraph are therefore used to identify which antennas are mostly suited for a given set of requirements related to the wireless application. For example, a signal bandwidth of 100 kHz would need an antenna bandwidth of at least 100 kHz, so that all the spectral components of the signal are equally treated, and signal distortion is kept to a minimum. The antenna radiation characteristics also play a significant role in wireless system design. A point-to-point link with direct line-of-sight between that transmitter and the receiver would probably need a high-gain antenna to maximize the power of the signal that reaches the receiver. In order for antenna designers to have a better view of the challenges involved when integrating an antenna into a wireless communication system, some basic communication concepts are described here, related to the wireless channel, and the MIMO concept.

2.3.1 *The Antenna Perspective of the Wireless Channel*

Once a wave departs from the antenna of a wireless transmitter, it enters the wilderness of the wireless channel. Antenna engineers are quite familiar with the equations describing wave propagation in vacuum or homogeneous media in an analytical form. However, the wireless channel is neither empty nor homogeneous. It includes a number of discontinuities which reflect, diffract, and scatter incident waves. The result is that any realistic environment thrives with multiple copies of the same transmitted signal. Whenever a receiver is placed within such an environment, it captures a vector sum of these wave components. This effect, as described in

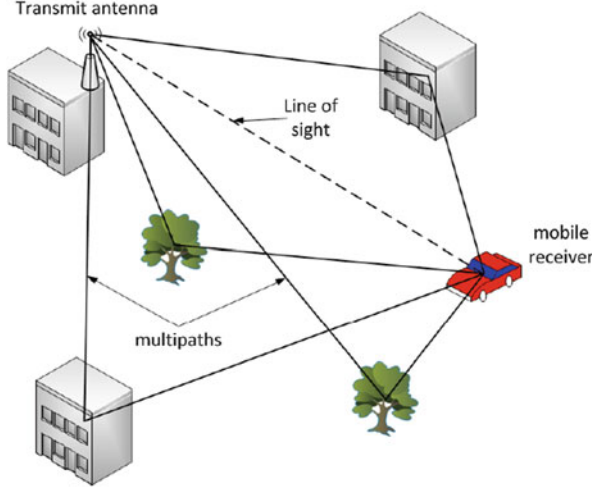


Fig. 2.11 The multipath environment of the wireless channel

Fig. 2.11, is called multipath, and taking into account the transmit and receive antenna patterns, it can be modeled as

$$h(\tau) = \mathcal{F}^{-1}[H(f)] = \sum_{i=1}^N P_T(\varphi_{T,i}) P_R(\varphi_{R,i}) g_i \delta(\tau - \tau_i) \quad (2.12)$$

In Eq. (2.12) shown above, the temporal dependence has been removed for convenience. Moreover, L is the number of multipath components (the number of different copies of the same signal that are added on the receiver end), $\varphi_{T,i}$ is the angle of departure (AoD) of the multipath component from the transmitter antenna, $\varphi_{R,i}$ is the AoA of the multipath component on the receive antenna, $P_T(\varphi)$, $P_R(\varphi)$, is the complex radiation pattern of the transmit and receive antennas, respectively, g_i is the complex gain (amplitude and phase) of the i th multipath component, and τ_i is the path delay. For simplicity, the metrics of equation (2.12) are shown to be time invariant, describing a linear, time invariant (LTI) system. Although these metrics are in general time-varying (the transceivers and objects in the wireless environment are in general not static), it is safe to assume for the analysis in this book that during transmissions the channel does not change dramatically. Since the channel can be represented as a LTI system, we can fully characterize it by its impulse response. This corresponds to the output of the system when the input is a Dirac (delta) function. A graphical representation of a typical wireless environment is shown in Fig. 2.12.

The impulse response of the channel gives to the wireless system designer useful information concerning the achievable rates of the channel. A general rule of thumb is that the larger the spread of the multipath delays, the larger the inter-symbol

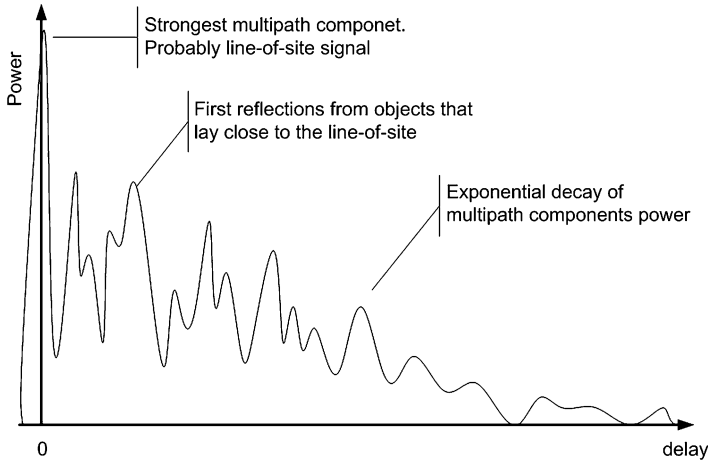


Fig. 2.12 A typical impulse response of a wireless channel

interference and consequently, the smaller the rate that the wireless channel can achieve. The second central moment of the power-delay profile of the channel is called the *rms delay spread* of the channel and it is derived directly from (2.12), according to

$$\sigma_\tau = \sqrt{\tilde{\tau} - (\bar{\tau})^2}, \quad \tilde{\tau} = \frac{\sum_i g_i^2 \tau_i^2}{\sum_i g_i^2}, \quad \bar{\tau} = \frac{\sum_i g_i^2 \tau_i}{\sum_i g_i^2} \quad (2.13)$$

If the delay spread of the channel is less than 10% of the communication symbol duration, then we may assume that the channel treats all frequency components of the signal equally, introducing minimal distortion to the signal form. In this case, we may represent the channel as a single complex number, which equally alters the amplitude and phase of all signal components. The channels that obey to the aforementioned criterion are called flat-fading channels and are the primary focus of this book. It is evident from (2.12) that the antenna radiation pattern plays a significant role in the channel impulse response. The multipath gains can be directly controlled by the antenna response vector toward the angles of arrival and departure. One straightforward advantage of this observation in wireless systems is that by using directional beam patterns, delayed multipath components may be suppressed to reduce the delay spread of the impulse response. This is achievable in cases where there is correlation between multipath delay times and AoD, AoA. This correlation can be determined by measurements and is illustrated in power-delay-angle profiles as the one shown in Fig. 2.13.

Therefore, with the use of appropriate antennas we may increase the rate of the wireless channel. Alternatively, we may control the individual feeds to the diverse

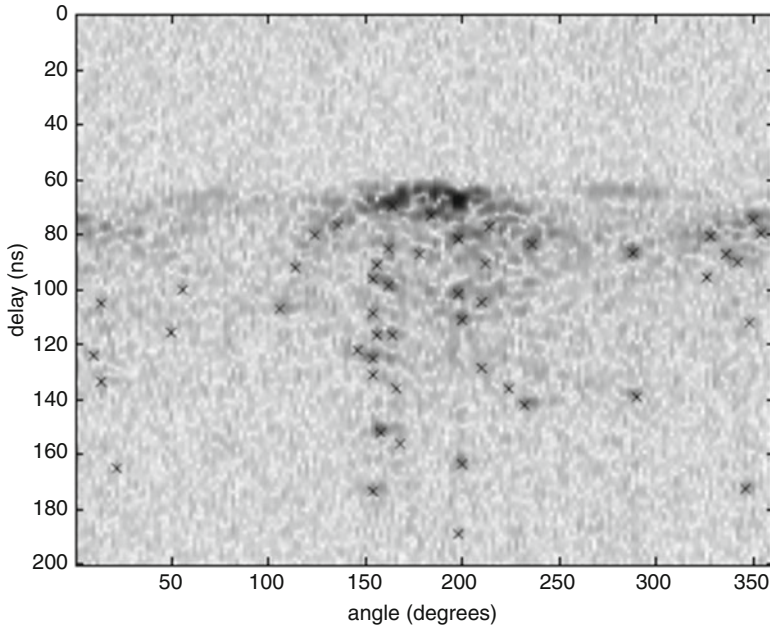


Fig. 2.13 Power delay angle profile of a wireless environment, copied from [7]

multipath components in such a way so as to maximize the received signal strength at the receiver. All the aforementioned advantages can be easily explained by taking into account only the radiation pattern of the antenna, with the help of (2.12). Therefore, from the antenna perspective, the channel characteristics can be processed by means of controlling the antenna radiation pattern, a method that is often referred to as beam-forming. However, there is a class of techniques that exploit the multipath nature of the wireless channel for increasing the capacity of links, which takes into account the structure of the antenna, assuming that multiple antenna elements are used both at the transmitter and receiver of the communication link. This class of algorithms which originated from the signal processing community, called MIMO algorithms, are described in the next paragraph.

2.3.2 MIMO Basics

MIMO stands for multiple input–multiple output systems. These are wireless communication systems that have multiple antenna elements both at the transmitter and the receiver of the communication link. Although in the signal processing community all the antenna elements used for communication are assumed to have isotropic radiation patterns, the exact form of the signals sent and received from

diverse directions in space is affected by the transmitter and receiver antenna patterns, respectively. This is illustrated in the expansion of Eq. (2.12), to include multiple antenna element systems:

$$\mathbf{H}(\tau) = \begin{bmatrix} h_{11}(\tau) & h_{12}(\tau) & \cdots & h_{1N}(\tau) \\ h_{21}(\tau) & h_{22}(\tau) & \cdots & h_{2N}(\tau) \\ \vdots & \vdots & \ddots & \vdots \\ h_{M1}(\tau) & h_{M2}(\tau) & \cdots & h_{MN}(\tau) \end{bmatrix} \quad (2.14)$$

In the case of flat fading wireless channels, (2.14) is reduced to

$$\mathbf{H} = \begin{bmatrix} h_{11} & h_{12} & \cdots & h_{1N} \\ h_{21} & h_{22} & \cdots & h_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ h_{M1} & h_{M2} & \cdots & h_{MN} \end{bmatrix} \quad (2.15)$$

which means that the signals on the antennas of the receiver will be

$$\mathbf{y} = \mathbf{H}\mathbf{x} + \mathbf{n} \quad (2.16)$$

This signal model can be used to determine the number of independent wireless channels that are created when multiple antennas are used at both ends of a communication link. In linear algebra, a common method to identify the number of independent equations out of system of multiple equations is the singular value decomposition. According to that, the system described by (2.16) can be rewritten as,

$$\mathbf{y} = \mathbf{U}\mathbf{S}\mathbf{V}^H \mathbf{x} + \mathbf{n} \quad (2.17)$$

where, $\mathbf{U}$ is a $M \times M$ complex unitary matrix, $\mathbf{S}$ is a $M \times N$ diagonal matrix, with nonnegative real numbers in descending order on the diagonal, and $\mathbf{V}$ is a $N \times N$ complex unitary matrix. The number of nonzero values of the diagonal matrix $\mathbf{S}$ corresponds to the number of independent channels created from the use of multiple antennas. Therefore, that maximum number of independent channels in the best case scenario, where all numbers in the $\mathbf{H}$ matrix are independent, is equal to the minimum value between N (the number of transmit antennas) and M (the number of receive antennas¹). In other words, if we want to maximize the number of independent channels of a MIMO link, the elements of the $\mathbf{H}$ matrix should be independent and identically distributed random variables.

¹It must be noted, however, that in realistic scenarios, measured singular values can never be exactly zero. In that case, we choose to use the independent channels that their singular value power is greater than the noise threshold.

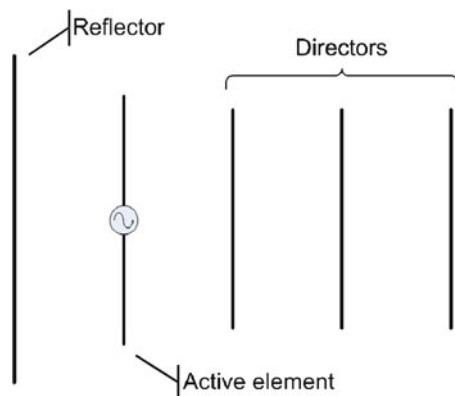
In real MIMO systems, there are quite a few methods for achieving this independence between the diverse channels seen by the different antenna elements. The most common method is to place antenna elements of transmit and receive arrays far from each other. At a distance approximately equal to half the wavelength of the carrier waveform the sums creating the elements of matrix $\mathbf{H}$ become uncorrelated. Therefore, if antenna elements are placed half a wavelength from each other and have no mutual coupling, they will act as virtually independent transceivers. It must be noted that antenna coupling is a different phenomenon from channel dependence, although they both can contribute to decreasing the number of independent channels that the multiple antenna system can create. A different method for creating independent channel instances in the channel matrix, $\mathbf{H}$, is to take advantage of the orthogonality between different polarizations of the electromagnetic waveform. As described in previous paragraphs, antennas that use different polarizations are not capable of efficiently communicating with each other. Furthermore, in the same environment, differently polarized signals experience different reflections and diffractions. In other words, waves of different polarizations “see” different channels in the same environments. This means that if multiple antennas on the transceivers have different polarizations, then the channel instances of the channel matrix will again be independent, leading to the maximization of independent channels seen in the same link.

Finally, as seen in Eq. (2.15), if each antenna element of the multi-element transceiver arrays has a different radiation pattern, then the elements of the channel matrix can again be engineered to be independent. This last observation has motivated the approach described in this book. Although the analysis of this approach is in the focus of the next chapter, it is safe to say here that the latter approach provides significant advantages to the implementation of MIMO arrays, related to the size and cost of their implementation. This has been achieved by changing our perspective to what has been previously described as a curse of multi-element arrays: mutual coupling.

2.4 The Blessing of Mutual Coupling

Although mutual coupling has been described as a curse for multi-element antenna arrays, there are quite a few cases in the history of antenna development that mutual coupling has been used as a design tool in order to meet certain antenna requirements. The most common example of this approach is the Yagi–Uda antenna. The Yagi–Uda antenna is a linear multi-element array, shown in Fig. 2.14, consisting of one active dipole or loop antenna and a number of parasitic dipoles. The parasitic dipoles are placed close to the active element, so that strong currents are induced to them. If the inter-element spacing between the elements of the array is carefully engineered, then the resulting current distribution of the array is such that a highly directional antenna array is formed.

Fig. 2.14 Structure of a Yagi–Uda array



The form of the currents on the parasitic elements of the array can be carefully designed by adjusting not only the inter-element spacing but also the length of the parasitic dipoles. Short dipoles have no impact on the radiation pattern of the active element, whereas dipoles with close-to-resonance-lengths produce currents of significant amplitude, therefore significantly affecting the radiation pattern of the whole structure. This property of parasitic arrays has been widely used to create simple smart antennas, called “switched parasitic arrays” (SPAs). These are based on a simple idea: since the currents on parasitic elements largely depend on whether these parasitics have resonant lengths or not, a simple switch can control the amplitude of the induced currents by simply changing the electric length of these elements. An example using dipoles is shown in Fig. 2.15.

When the switch in Fig. 2.15 is in the OFF state, then the two branches of the parasitic dipole are open-circuited; therefore the parasitic element does not have a resonant length and the amount of current induced to it is minimal. Therefore, the radiation pattern created by the structure is dominated by the radiation pattern of the single active element, which in this case is the omni-directional pattern of the dipole antenna. On the other hand, when the switch is in the ON state, the two branches of the parasitic dipole are short-circuited, the parasitic element acquires a resonant length, and strong currents are induced to it. Therefore, the radiation pattern of the structure changes to the directional pattern shown in Fig. 2.15.

The property described above has been widely used to create simple smart antenna structures, which can steer a directional beam toward a finite number of angular directions. The most common antenna structure used for this purpose is the circular array. The latter has a single active element placed at the center of the structure, and a number of parasitic elements evenly spaced on the periphery of a circle with radius d , around the active element. This structure, shown in Fig. 2.15, can create a number of diverse radiation patterns, by simply open-circuiting and close-circuiting the parasitic elements’ input.

SPAs were a significant finding in the antenna design field. They enabled engineers to manufacture smart antenna arrays with two main advantages over their active element counterparts. SPAs needed only a single RF feed to operate and a

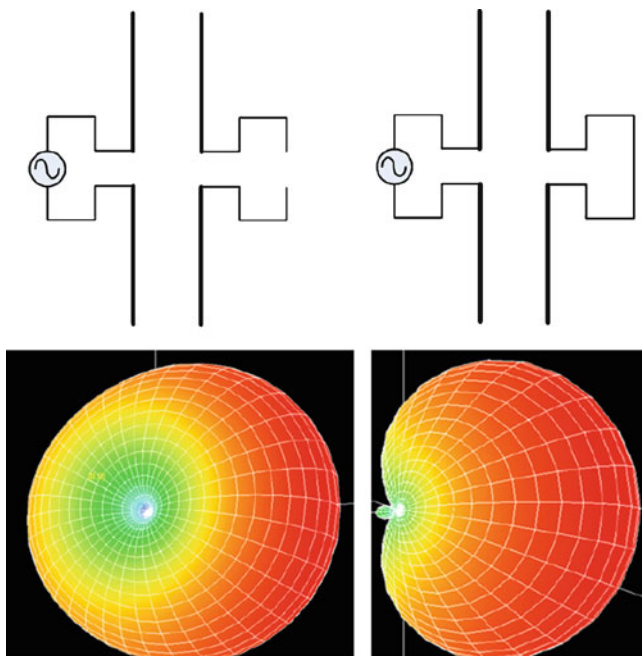


Fig. 2.15 A typical switched parasitic array structure and radiation patterns

number of digital control lines. This meant that smart antenna functionality could be supported with minimal additional cost and complexity, compared to single antenna architectures. Furthermore, since mutual coupling of parasitic arrays does not have to be minimized, and parasitic elements do not have to be matched to the specific impedance of an RF circuit, there are actually less constraints to the design SPA-based smart antennas. For the aforementioned reasons, SPAs have been widely used in commercial wireless communication systems and have found their way even to handheld wireless devices (e.g., Ericsson, HTC smartphones).

However, there is a price to pay for these advantages in terms of the simplicity and cost of SPAs. SPAs cannot control their radiation pattern to the extent that their active array counterparts can. The latter have the capability to create infinite combinations of signals at the ports of the antenna elements of the array. SPAs on the other hand can only create a finite number of pre-defined patterns. Although these predefined patterns can be adequate for a number of applications, beam switching is not as powerful when compared to space-time processing algorithms which active element arrays can support. Should we want to combine the powerful capabilities of active element arrays with the low-complexity and low-cost design characteristics of SPAs, we would have to search back to the times where DSP was neither powerful nor widespread. Namely, in the mid-1970s, Luzwick and Harrington [8] presented an idea of how to control the currents of parasitic elements directly at the antenna feeding network, without the need for additional RF front ends. The idea can be

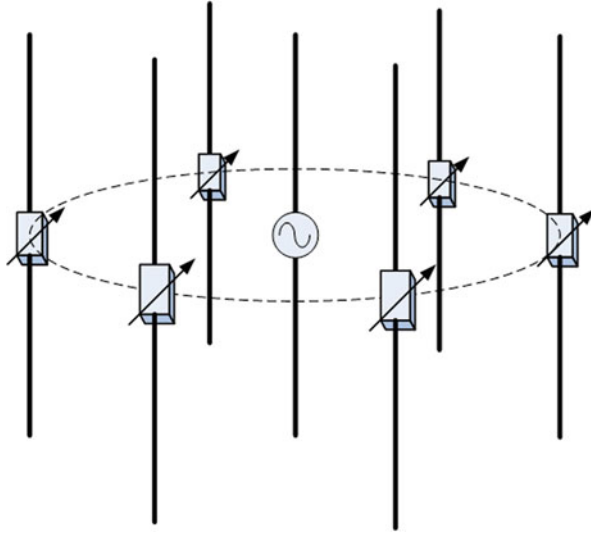


Fig. 2.16 The concept of the ESPAR antenna

thought of as an extension of SPAs. The difference is that the control circuit does not only have two switching states (ON and OFF), but it can realize a wide range of load impedances. For each different load of the parasitic, the form of the currents induced to it is also different, thus leading to more accurate control of the antenna radiation characteristics. This reconfigurable load circuit can be as simple as a single varactor, controlling the imaginary part of the parasitic load impedance. This basic idea was revisited in 2,000 in ATR laboratories, Japan, by a group led by Gyoda and Ohira [9]. They focused on the circular array structure, having a number of parasitic elements on the periphery of the circle and a single active element placed at the center of the circle, as shown in Fig. 2.16. They termed this design as “Electronically Steerable Passive Array Radiator,” or ESPAR antenna.

The ports of the parasitic elements are short-circuited and connected to variable reactor (varactor) diodes, which can change their capacitance according to a control voltage. Therefore, the parasitic loads have load impedances with a real part equal to zero and an imaginary part which can be directly controlled by a DSP. Due to the strong mutual electromagnetic coupling among all elements, currents are induced to all parasitics. Such currents depend on the array’s geometry, i.e., element arrangement and inter-element distance. Furthermore, by tuning the varactor values, henceforth called weights, the effective coupling among all neighbor elements is changed causing a corresponding change at all currents and consequently at the radiation pattern:

$$P(\phi) = \mathbf{i}^T \mathbf{a}(\phi) = \sum_{n=0}^{M-1} i_n a_n(\phi) \quad (2.18)$$

where M is the number of ESPAR elements and $a(\phi)$ is the steering vector of the array. The $(1 \times M)$ current vector i is given by $\mathbf{i} = (\mathbf{Z} + \mathbf{X})^{-1} \mathbf{v}$ where $\mathbf{Z}$ is the $(M_{esp} \times M_{esp})$ electromagnetic coupling matrix of the ESPAR antenna, $\mathbf{X} = \text{diag}[50 \text{ j}x_1 \dots \text{j}x_{M_{esp}-1}]$ is the load diagonal matrix that adjusts the radiation pattern, while $\mathbf{v} = [1 \ 0 \ 0 \ \dots]^T$ is a $(M_{esp} \times 1)$ vector.

In the initial paper of Gyoda and Ohira [9], a seven-element circular structure was considered, with one active and six parasitic elements, placed at the periphery of a circle with radius $\lambda/4$. In this symmetrical structure, the impedance matrix of the array is given by (2.11).

It is evident that due to the symmetrical nature of the structure, most of the impedance values of the matrix are equal. However, one can think of several non-symmetrical structures that could fall within the same concept. This basic structure has been used to demonstrate that several smart antenna algorithms (beam-forming [10], diversity [11], AoA estimation using ESPRIT [4, 5] or MUSIC [2, 3]) can be implemented using a single RF front end. However, until recently this architecture could not be used for achieving the benefits of MIMO communications. Integrating single RF MIMO technology in applications with quite strict size and cost constraints, e.g., mobile devices, would significantly increase the performance of wireless communications, and it is the main focus of this book. In order to achieve this goal, it is required that the two diverse worlds of the smart antenna divide are brought together.

2.5 Bridging Two Diverse Worlds

In this chapter we tried to explain some fundamental concepts related to the antenna perspective and the digital signal processing perspective of building smart antenna systems. These diverse worlds have contributed significantly to the development of smart antennas. However, in this era of personal mobile communications, the increasing demand for handheld, lightweight, power efficient, and cost-effective wireless devices, brings the aforementioned approaches to their limits. Further development of smart antenna systems now calls for a broader view of the problem, bringing together knowledge from two diverse worlds: antenna design and signal processing.

In the previous paragraph we have presented an example of such an approach, showing how smart antenna systems can be implemented using a single RF front end, by taking advantage of mutual coupling. In other words, by bridging the smart antenna divide, we have taken an antenna property that has for long been considered as a flaw of antenna arrays by the signal processing world, and turned it into a key design tool for delivering high-end smart antennas at the cost SPAs. Thus, parasitic arrays have already proved their capabilities for implementing several smart antenna algorithms like beam-forming, diversity, and AoA estimation. However, one of the most promising technologies for increasing the performance of wireless communications, namely MIMO transmission, seems

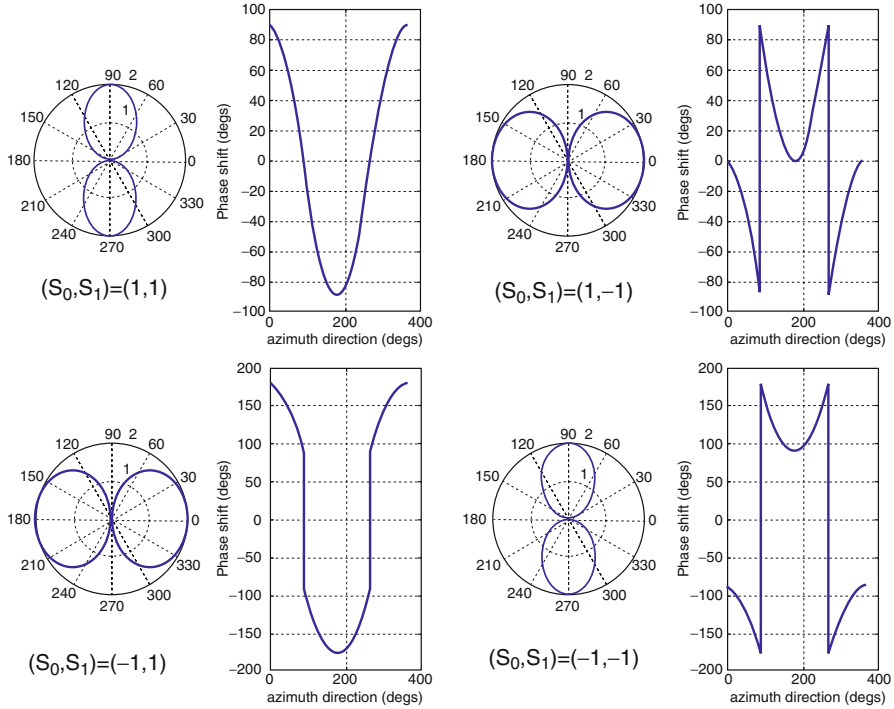


Fig. 2.17 Radiation patterns (amplitude component on the *left* and phase component on the *right*) created during different symbol periods using a traditional two-element antenna array with inter-element spacing equal to $\lambda/2$. The symbol streams are modulated using two BPSK signals, s_0 and s_1

irrelevant to single input parasitic arrays. After all, MIMO stands for multiple input–multiple output. Nevertheless, by closer inspection, the relationship of parasitic arrays to conventional MIMO would reveal fundamental similarities despite the different approaches. Take for example the case of a conventional two-element MIMO system that uses two dipole elements at a distance of $\lambda/2$ from each other. If the MIMO array is driven by two uncorrelated binary-phase-shift-keyed (BPSK) symbol streams, then on each symbol period different radiation patterns would appear at the far-field of the antenna array, as shown in Fig. 2.17.

BPSK modulation codes symbols as signal of the same frequency, but opposite phase. When these symbols are fed to the two diverse antennas of the MIMO array, then the inputs of the antenna elements will be either co-phased or opposite-phased. These two possible states produce the two radiation patterns of Fig. 2.17, when the distance between the elements is equal to $\lambda/2$. Therefore, MIMO arrays can be seen from the antenna perspective as beam-forming arrays that change their radiation pattern on each symbol period [12]. Since beam-forming arrays can be implemented using parasitic antennas, why is it not possible that the latter can also be used for implementing MIMO systems? Naturally, this different approach would

need new methods and tools for modeling the antenna array within the wireless environment. The main approach for bringing together the diverse worlds of antenna design and signal processing for creating MIMO systems is termed “wave-vector domain analysis” and is presented in the next chapter.

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