Diplomarbeit

Filter Design for the Spectral Optimization of UWB Signals

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Erklärung

Hiermit versichere ich, dass ich die vorliegende Arbeit selbständig und ohne unerlaubte Hilfsmittel angefertigt habe.

Karlsruhe, den 20.10.2008

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1 Introduction to Ultra-Wideband

Ultra-wideband (UWB) is a radio technology that uses a large portion of the spectrum but emitting very low energy levels for short-range communications. UWB transmission systems are characterized by using an instantaneous bandwidth greater than 500 MHz or a fractional bandwidth of more than 20%. The fractional bandwidth is defined as $B/f_c$, where $B = f_H - f_L$ denotes the $-10$ dB bandwidth and center frequency $f_c = (f_H + f_L)/2$ with $f_H$ being the upper frequency of the $-10$ dB emission point, and $f_L$ the lower frequency of the $-10$ dB emission point. UWB systems with $f_c > 2.5$ GHz need to have a $-10$ dB bandwidth of at least 500 MHz, while UWB systems with $f_c < 2.5$ GHz need to have fractional bandwidth at least 0.20.

UWB uses base band pulses that do not require IF processing of ultra-short duration, typically about a nanosecond, and spreads the signal energy very thinly across the entire bandwidth being used.

1.1 Advantages

These characteristics make Ultra-wideband radios be very appreciated for the advantages they provide. Such advantages are enumerated and described next:

- Very high data rates: The transmission of about 1 or 2 giga-pulses per second enables short-range gigabit-per-second communications [1]. According to Shannon’s information transmission capacity expression\(^1\), the channel capacity increases linearly with frequency bandwidth, and decreases logarithmically with the signal to noise ratio. Since UWB has wide frequency bandwidth, it is suitable for high data rate communications.

- Coexistence: Intrinsic low-power transmission from Ultra-Wideband helps mitigate interference [2], [3]. This permits the use of several Ultra-Wideband devices adequately separated at the same time.

- Ultra high precision ranging: The use of extremely short pulses facilitates the obtaining of the time of arrival of the signal and hence estimating the position of the target [4].

- Multipath immunity: While in narrowband radio systems the continuous carriers are susceptible to destructive interference due to multipath in environments having reflecting objects, in the UWB radio system the multipath signal energy arrives

\(^1\) $C = B \log(1 + S/N)$ where $C$ is the channel capacity in bits per second if the logarithm is taken in base 2, $B$ is the channel bandwidth in hertz and $S$ and $N$ are the signal and noise powers measured in watts
at a different time than the direct path energy and cannot suffer destructive interference since short pulses rarely overlap [5]. In addition, a multiple correlator system can benefit from this multipath energy to improve the performance of the system.

- Low consume: The regulated power levels are very low [6], (below $-41.3$ dBm) which entails low consume as a radio wave emitter.

- Security: As the emitted power density of Ultra-Wideband signals is very low (usually below environment noise) only a receiver that knows the schedule of the transmitter can decode the information. Furthermore, narrow band receivers cannot distinguish an Ultra-Wideband signal from the environment noise [7].

- Simplicity and low cost: Ultra-Wideband uses carrier-less radio impulses [8], so it only has base-band processing and it does not need intermediate frequency processing. This makes Ultra-Wideband devices much cheaper than other communication systems. In the super-heterodyne architecture, the base-band signal is first converted to intermediate frequency signal by multiplexing with a local oscillator frequency, afterwards this intermediate frequency signal is converted to a higher frequency signal which is called radio frequency signal. However, in Ultra-Wideband systems, a local oscillator is not needed, no up- or down-converters are necessary, hence, they are simple and of low cost.

- Enhanced capability to penetrate through obstacles [9].

### 1.2 Applications

All these characteristics open the door for many interesting applications in several fields. Ultra-Wideband transceivers can send and receive high-speed data at short distances. Nowadays, Ultra-Wideband systems are aiming at indoor applications within several meters at bit rates up to hundreds of megabits per second. This could be useful for replacing cables (connect digital cameras to a computer, wire-less USB, etc) and for high-speed wireless local area network communications [10], [11]. Moreover, the addition of an Ultra-Wideband transceiver into a handheld device would barely modify its size, weight or shape since the CMOS technique allows reduced dimensions [12], [13].

Ultra-Wideband systems may also be used to determine the separation distance between two objects due to its high precise time resolution. The use of very short pulses avoid that the reflected pulses overlap, making them resolvable. Ultra-Wideband can also be utilized for medical purposes as it can propagate through the human body as well. An Ultra-Wideband radar is able to detect, non-invasively, the movements of the heart wall [9], for example. In addition, taking advantage of its capability to penetrate through obstacles, Ultra-Wideband systems may be used to detect and image the objects inside enclosed spaces, behind walls or under the ground. This allows to sense the presence of pipes or land mines or to image the positions of people inside the rooms by
placing the Ultra-Wideband radar outside the house. So it might be useful as a victim search device in emergency situations.

Another researching area is in the automobile field [14]. Ultra-Wideband vehicular radars measure the location and movements of objects around a vehicle by transmitting pulses and analysing the reflected signals. This makes auto navigation or collision avoidance possible, just to mention some examples.

### 1.3 Regulations

#### 1.3.1 United States

In 2002, the Federal Communications Commission (FCC) set the rules [6] to permit the operation of devices incorporating the Ultra-Wideband technology. The standards were established so that Ultra-Wideband devices can operate using spectrum occupied by existing radio services without causing interference. Such thing is achieved by operating below the Part 15 limit\(^2\).

As shown in Figure 1.1, for indoor applications, the average emission limits for Ultra-Wideband devices are \(-41.3\) dBm/MHz from 0 Hz to 0.96 GHz, \(-75.3\) dBm/MHz from 0.96 GHz to 1.61 GHz, \(-53.3\) dBm/MHz from 1.61 GHz to 1.99 GHz, \(-51.3\) dBm/MHz from 1.99 GHz to 3.1 GHz, \(-41.3\) dBm/MHz from 3.1 GHz to 10.6 GHz and \(-51.3\) dBm/MHz from 10.6 GHz above.

![Emission limits for Ultra-Wideband indoor applications in the United States.](image)

The emission limits are defined as the equivalent isotropically radiated power (EIRP) in dBm\(^3\) per 1 MHz bandwidth (dBm/MHz). Nevertheless, it is the electric field (µV/m) that causes interference to radio communications and a given electric field strength does

\(^2\)FCC Part 15 rules [15] cover both unintentional radiators (devices such as computers which may generate radio signals but are not intended to transmit them); and intentional radiators (such as garage door openers, cordless telephones, etc.).

\(^3\)dBm deciBel relative to one milliwatt
not directly correspond to a certain level of transmitter power. The relation between power and field strength can be approximate with:

\[
\frac{PG}{4\pi R^2} = \frac{E^2}{\eta}
\]

(1.1)

where:

- \(P\) is the transmitter power in Watts;
- \(G\) is the gain of the transmitting antenna;
- \(R\) is the radius in meters of the sphere at which the field strength is measured;
- \(E\) is the field strength in volts/meter; and,
- \(\eta\) is the characteristic impedance, which is \(120\pi\ \Omega\) in free space.

### 1.3.2 Europe

In 2007, the European Commission (EC) issued details of the licensing regulations for Ultra-Wideband networking in Europe [16]. The EC has chosen to make use of only part of the spectrum that was approved for use in the US in 2002. In Europe, the equivalent isotropic radiated power (EIRP) value of \(-41.3\ \text{dBm/MHz}\) is applied over the 6.0 to 8.5 GHz. This narrower band is to provide more protection to other radio users while staying close to the US regulations. The European Commission regulations are plotted in Figure 1.2.

![Bild 1.2: Emission limits for Ultra-Wideband indoor applications in Europe.](image-url)
2 Motivation and goals of this work

As seen in the previous chapter, Ultra-Wideband systems have power spectral density constraints set by both the Federal Communications Commission in United States and the European Commission that have to be accomplished. These constraints define a mask which cannot be exceeded in order to avoid interfere other radio devices. This, together with the fact that Ultra-Wideband systems are forced to operate at very low power density levels (below $-41.3$ dBm/MHz), makes critically important not only respecting the mask but also filling it up as much as possible. That means, taking best advantage of the available resources. In this work, a solution for this problem is studied, realized and tested. All the elements that should be taken into consideration are described next.

2.1 Description of the initial scenario

Let consider the relevant block diagram of a Ultra-Wideband transmission system in Figure 2.1. The pulse generator [17] creates a pulse waveform in response to each data value, then it is radiated to the air by the transmitting antenna. This on-the-air spectrum must accomplish the emission regulations set by the competent commission.

\[ G(f) \quad A(f) \quad S(f) \]

Bild 2.1: UWB transmission system.

In the previous figure, $G(f)$ is the pulse spectrum of a given pulse generator, $A(f)$ is the antenna transfer function [18], [20], [21] and $S(f)$ is the on-the-air spectrum of the transmitted signal. In contrast to conventional narrow-band systems, where the waveform distortion by the antenna is negligible, in Ultra-Wideband systems there is an important signal distortion by Ultra-Wideband antennas which has to be taken into consideration. In such a wide bandwidth, the behavior of the antenna depends strongly on the frequency. Therefore, the antenna structure acts like a filter with its corresponding transfer function. It is also important to emphasize that the transmitted signal is mathematically expressed as the first time derivative of the signal waveform applied to the transmitting antenna [18], [22]. Now, it is clear that antennas have a severe influence on the radiated signal and they cannot be characterized as a mere differentiator in time.
2.2 Possible solutions

By way of example, we will consider that the pulse spectrum of our pulse generator is the one depicted in Figure 2.2, which is emitted at maximum allowed power to exploit, as far as possible, all the available spectrum. We also have at our disposal the idealized version of the transfer function of the emitting antenna (Figure 2.3) where the $j2\pi f$ factor, that represents the first time derivative in the frequency domain, has not been included. Consequently, to compute the expression of the radiated, is necessary to add it in the equation. In this first case, the antenna provides constant gain within all the frequency range, which means that it can be considered as differentiator in time. With all this, the radiated spectrum (Figure 2.4) by the antenna will be:

$$G(f) \cdot A(f) \cdot j2\pi f = S(f)$$

By observing the spectrum $S(f)$ in Figure 2.4, emitted by the ideal system, we realize that it violates the mask (FCC mask in this example) significantly at low frequencies, while not making an efficient use of the available spectrum from around 6 GHz on. A solution to alleviate this problem will be tackled in the next section.

2.2 Possible solutions

Many solutions have been proposed to solve this problem [23, 24, 25, 26]. Most of these solutions imply using certain pulses that are somehow difficult to implement or entail developing digital systems. The scope of this thesis is to find a method to achieve near-maximum transmitted energy while satisfying the specified emission limits, that is, to maximize the transmitted pulse energy subject to the mask constraint, utilizing analog circuits and without modifying the elements of the transmitting end of our Ultra-Wideband link. These elements are the pulse that our generator provides and the transmitting antenna. With the knowledge of the created pulse by the pulse generator, the antenna transfer function and the desired on-the-air pulse spectrum it is possible to obtain a solution for the aforementioned challenge.

2.3 Proposed solution

The solution that is proposed in this work is show in Figure 2.5. The idea is to add a microwave passive circuit in between the pulse generator and the transmitting antenna
2.4 AVAILABLE DATA

In order to compute the expression in (2.3), it is necessary to have all the factors involved in it. Firstly, the waveform generated by our pulse generator has been measured in the time domain and is plotted in Figure 2.6. However, this data is still useless to make a pre-distortion of the signal so that the desired spectrum is radiated. As seen in Figure 2.5, a pulse is generated in response to each date value. Then, a filter shapes that pulse so that the antenna output has the desired form. According to the block diagram in the frequency domain, the expression in (2.2) can be written.

\[ G(f) \cdot H(f) \cdot A(f) \cdot 2\pi f = S(f) \]  (2.2)

where:
- \( G(f) \) - pulse spectrum created by our pulse generator
- \( H(f) \) - required shaping filter response that is to be designed
- \( A(f) \) - transfer function of the transmitting antenna
- \( S(f) \) - desired pulse spectrum at the antenna output

The \( 2\pi f \) factor expresses, in the frequency domain, the time differentiation that the antenna performs. Then, solving (2.2) for \( H(f) \) we obtain (2.3).

\[ H(f) = \frac{S(f)}{G(f) \cdot A(f) \cdot 2\pi f} \]  (2.3)

At this point, it only remains to define what desired pulse spectrum at the antenna output (\( S(f) \)) means. The aim is to obtain a spectrum emitted by the transmitting antenna that matches, as best as possible, a certain mask while not exceeding it in any way. In this work, the two masks defined in section 1.3, will be used to compute the expression of the desired shaping filter. The ideal case is to take \( S(f) = S_{\text{mask}}(f) \), where \( S_{\text{mask}}(f) \) is either the spectrum plotted in Figure 1.1 or in Figure 1.2. For our purposes it will be sufficient to consider a flat spectrum at the antenna output within a certain frequency range. For the case of the FCC mask, the frequency range will be from 3.1 GHz to 10.6 GHz and for the EC mask, the frequency interval will be from 6 GHz to 8.5 GHz.

2.4 Available data

In order to compute the expression in (2.3), it is necessary to have all the factors involved in it. Firstly, the waveform generated by our pulse generator has been measured in the time domain and is plotted in Figure 2.6. However, this data is still useless to
compute (2.3) since everything is needed in the frequency domain. By applying the fast Fourier transform (FFT) on the time domain waveform, the spectrum in Figure 2.7 is obtained. It is observed that the FFT of the pulse returns a very rough curve. Therefore, a smoothed version of this curve, which can be seen together with the direct FFT, will be used. The desired on-the-air spectrum $S(f)$, as it has been mentioned in the previous section, will be flat from 3.1 GHz to 10.6 GHz for the FCC mask and from 6 GHz to 8.5 GHz for the EC mask. Finally, the transfer function of the antenna is required to have all the elements in the expression.

In this work, a solution for two different planar antennas is proposed. The Vivaldi antenna and the Bow-tie antenna (Figure 2.12) whose transfer functions have been measured and are plotted in Figure 2.8 and Figure 2.9 respectively. It is worth mentioning that these antenna do not radiate below 2 GHz.
2.5 Computed ideal filters

As we can note by observing the figures, the transfer function of the Vivaldi antenna can be approximated with a straight line with negative slope. Unlike the Bow-tie antenna, whose behaviour is quasi-constant within the 3.1 to 10.6 GHz band. This flat approximation is much closer to the real transfer function in the European frequency band, that is from 6 GHz to 8.5 GHz due to the narrower bandwidth. This is illustrated in Figures 2.10 and 2.11 where a closer view of the transfer function for the Bow-tie antenna is shown for both frequency ranges.

Bild 2.10: Measured transfer function of the Bow-tie antenna in the main beam direction, co-polarized component.

Bild 2.11: Measured transfer function of the Bow-tie antenna in the main beam direction, co-polarized component.

Bild 2.12: Bow-tie antenna (left) and Vivaldi antenna (right).

2.5 Computed ideal filters

Once we have reached this point, we can calculate the expression of the filter that is needed to obtain the desired spectrum at the antenna output. All needed data have been already provided in the previous section. The proposed solution of section 2.3 will be applied four times. That is, a shaping filter $H(f)$ to pre-distort the signal will be designed for the two different antennas, the Vivaldi and the Bow-tie antenna (Figure
2.5. COMPUTED IDEAL FILTERS

2.12), and for the two masks described in previous sections: the FCC mask and the EC mask. Using the expression (2.3), the shape of the filters $H(f)$ for all four cases is found and plotted. In Figure 2.14 the ideal filter response ($S_{12}$ parameter) for the FCC mask and the Vivaldi antenna is shown, in Figure 2.13 for the FCC mask and the Bow-tie antenna, in Figure 2.15 for the EC mask and the Vivaldi antenna, and finally in Figure 2.16 for the EC mask and the Bow-tie antenna. In order to compute these ideal $S_{21}$

![Bild 2.13: Vivaldi and FCC.](image1)

![Bild 2.14: Bow-tie and FCC.](image2)

parameters, the linear approximation of each antenna and the smoothed version of the pulse generated by the pulse generator have been used.

![Bild 2.15: Vivaldi and EU.](image3)

![Bild 2.16: Bow-tie and EU.](image4)

Obviously, the responses of the shaping filters that pre-distort the signal so that the output is flat in the FCC frequency range, are wider than the response of filters for the EC mask. This will be a key point in the design of the filters since the difference in the bandwidth is considerable and approximations that can be assumed in bandwidths from 6 GHz to 8.5 GHz are no longer valid in the frequency range from 3.1 GHz to 10.6
GHz. The other main difference between filter responses is the difference of attenuation levels in the two frequency edges. For the FCC case, the difference is greater than for the EC case due to the fact that the effect of the antenna is more evident because of the wider frequency interval. Moreover, the difference is bigger for the case of the Vivaldi antenna since the attenuation increases with the frequency and the shaping filter has to compensate it. These two main differences will be decisive in the filter design addressed in chapter 6.

In the next chapters, basic theory to understand the forthcoming filter design chapter will be presented since different microwave structures will be designed and simulated [27] to approximate as far as possible these ideal filter responses. They will also be constructed in order to perform measurements and test their correct behaviour in practice. Finally, a time-domain analysis will be done to complete the thesis.
3 Basic concepts of filters

The transfer function of a filter is the mathematical expression of the $S_{21}$ parameter:

$$|S_{21}(j\Omega)|^2 = \frac{1}{1 + \epsilon^2 F_n^2(\Omega)} \quad (3.1)$$

Where $\epsilon$ defines the ripple, $F_n(\Omega)$ is the filtering function of degree $n$, and $\Omega$ is a frequency variable, that represents the radian frequency variable of a lowpass prototype filter whose cutoff frequency is at $\Omega = \Omega_c$ for $\Omega_c = 1$ rad/s.

According to their frequency response, four kind of filters can be found:

- **Bandpass filters**: select only a desired band of frequencies.
- **Bandstop filters**: eliminate an undesired band of frequencies.
- **Lowpass filters**: allow only frequencies below a cutoff frequency to pass
- **Highpass filters**: allow only frequencies above a cutoff frequency to pass

### 3.1 Lowpass prototype filter

The lowpass prototype filter is useful in filter design because it can be used as template to produce a modified filter design for a particular application. Values are normalized to make the source resistance equal to one, denoted by $g_0 = 1$, and the cutoff angular frequency to be unity, denoted by $\Omega_c = 1$ rad/s. Figure 3.1 shows a possible form of a lowpass prototype filter response. It might be either Butterworth or Chebyshev of order $n$, being $n$ even. Figure 3.2 illustrates the case of $n$ being odd. However, the form of the lowpass prototype filters are not unique. Figure 3.3 demonstrates a network structure to realize the same lowpass prototype filter as in Figure 3.1. Both configurations can be used since they provide the same filter responses. If $n$ is odd, the last reactive element is a series inductance instead of a shunt capacitance, as shown in Figure 3.4. It can be noted that $n$ corresponds to the number of reactive elements. The values $g_i$ may have different interpretations depending on its placing. That is, if $g_1$ accounts for the shunt capacitance, then $g_0$ is resistance while if $g_1$ accounts for the series inductance, $g_0$ is defined as conductance. The same occurs for $g_n$ and $g_{n+1}$. This way, the desired filter can be scaled or transformed from this nondimensionalised design. The utility of a prototype filter comes from the property that all kind of filters (lowpass, highpass, bandpass and stopband) with different bandwidths, frequencies and impedances can be derived from it by applying a scaling factor to the components of the prototype. To change from the frequency domain $\Omega$, where the lowpass prototype is defined, to a practical frequency domain $\omega$, a frequency transformation is needed. As explained above,
the source impedance is also normalized. Consequently, it is also required to perform
the following impedance scaling:

\[ \gamma_0 = \begin{cases} 
Z_0/g_0 & \text{for } g_0 \text{ being the resistance} \\
g_0/Y_0 & \text{for } g_0 \text{ being the conductance} 
\end{cases} \] (3.2)

### 3.1.1 Lowpass transformation

The lowpass transformation is used to transform the lowpass prototype to a lowpass
filter that can be implemented practically. This transformation is given by

\[ \Omega = \left( \frac{\Omega_c}{\omega_c} \right) \omega \] (3.3)
3.1. LOWPASS PROTOTYPE FILTER

where $\omega_c$ is the cutoff frequency. Using (3.3) along with the impedance scaling, the following element transformation results:

$$L = \left( \frac{\Omega_c}{\omega_c} \right) \gamma_o g \quad \text{for } g \text{ representing the inductance}$$

$$C = \left( \frac{\Omega_c}{\omega_c} \right) \frac{g}{\gamma_o} \quad \text{for } g \text{ representing the capacitance}$$

\[ \text{3.1.2 Highpass transformation} \]

For highpass filters whose desired cutoff frequency is $\omega_c$, the frequency transformation is

$$\Omega = -\frac{\omega_c \Omega_c}{\omega}$$

\[ \text{3.1.3 Lowpass prototype filter.} \]

\[ \text{3.1.4 Modification if } n \text{ is odd.} \]
which means that a reactive element $g$ will be mapped to:

$$j \Omega g \rightarrow \frac{\omega_c \Omega_c g}{j \omega}$$

Then, an inductive element in the lowpass prototype will be inversely transformed to a capacitive element in the highpass filter. And a capacitive element will be transformed to a inductive element. The final element transformation is given by

$$C = \left( \frac{1}{\omega_c \Omega_c} \right) \frac{1}{\gamma_0 g} \quad \text{for } g \text{ representing the inductance}$$ (3.7)

$$L = \left( \frac{1}{\omega_c \Omega_c} \right) \frac{\gamma_0}{g} \quad \text{for } g \text{ representing the capacitance}$$ (3.8)

### 3.1.3 Bandpass transformation

If the lowpass prototype filter needs to be transformed to a bandpass filter with a passband $\omega_2 - \omega_1$, where $\omega_1$ and $\omega_2$ are the passband-edge angular frequency, the required frequency mapping is

$$\Omega = \frac{\Omega_c}{FBW} \left( \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \right)$$ (3.9)

with

$$FBW = \frac{\omega_2 - \omega_1}{\omega_0} \quad \omega_0 = \sqrt{\omega_1 \omega_2}$$ (3.10)

where $\omega_0$ represents the center angular frequency and $FBW$ is defined as the fractional bandwidth. Following the same procedure as before and applying this transformation to a reactive element $g$ of the lowpass prototype, we obtain

$$j \Omega g \rightarrow j \omega \frac{\Omega_c g}{FBW \omega_0} + \frac{1}{j \omega FBW}$$

which implies that an inductive element $g$ in the lowpass prototype is transformed to a series $LC$ resonant circuit in the bandpass filter, while a capacitive element will transform to a parallel $LC$ resonant circuit. The element for the series $LC$ resonator are

$$L_s = \left( \frac{\Omega_c}{FBW \omega_0} \right) \gamma_0 g \quad \text{for } g \text{ representing the inductance}$$

$$C_s = \left( \frac{FBW}{\omega_0 \Omega_c} \right) \frac{1}{\gamma_0 g}$$

Similarly, the elements for the parallel $LC$ resonator in the bandpass filter are

$$C_p = \left( \frac{\Omega_c}{FBW \omega_0} \right) \frac{\omega_0}{g} \quad \text{for } g \text{ representing the capacitance}$$

$$L_p = \left( \frac{FBW}{\omega_0 \Omega_c} \right) \frac{\omega_0}{\gamma_0 g}$$
3.2. CHEBYSHEV RESPONSE

3.1.4 Bandstop transformation

Finally, the transformation to change from lowpass prototype to bandstop is the following

\[ \Omega = \frac{\Omega_c FBW}{(\omega_0 / \omega - \omega / \omega_0)} \]  

\[ \omega_0 = \sqrt{\omega_1 \omega_2} \quad FBW = \frac{\omega_2 - \omega_1}{\omega_0} \]  

(3.11)  

(3.12)

where \( \omega_2 - \omega_1 \) is the bandwidth to filter out. In this case, an inductive element \( g \) in the lowpass prototype is transformed to a parallel \( LC \) resonant circuit in the bandstop filter. A capacitive element \( g \) is transformed to a series \( LC \) circuit. These elements are:

\[ C_p = \left( \frac{1}{FBW \omega_0 L} \right) \frac{1}{\gamma_0 g} \quad \text{for } g \text{ representing the inductance} \]  

(3.13)

\[ L_p = \left( \frac{\Omega_c FBW}{\omega_0} \right) \gamma_0 g \]  

\[ L_s = \left( \frac{1}{FBW \omega_0 L} \right) \frac{\gamma_0}{g} \quad \text{for } g \text{ representing the capacitance} \]  

(3.14)

3.2 Chebyshev response

A Chebyshev filter exhibits equal-ripple passband and maximally flat stopband. The transfer function for a lowpass prototype Chebyshev filter is:

\[ |S_{21}(j\Omega)|^2 = \frac{1}{1 + \epsilon^2 T_n^2(\Omega)} \]  

(3.15)

where the ripple constant \( \epsilon \) is:

\[ \epsilon = \sqrt{\frac{L_{Ar}}{10^{\frac{L_{Ar}}{10}} - 1}} \]  

(3.16)

\( L_{Ar} \) being the passband ripple in dB and \( n \) the filter order. \( T_n(\Omega) \) is a Chebyshev function of the first kind of order \( n \), which is defined as

\[ T_n(\Omega) = \left\{ \begin{array}{ll} \cos(n \cos^{-1} \Omega) & |\Omega| \leq 1 \\ \cosh(n \cosh^{-1} \Omega) & |\Omega| \geq 1 \end{array} \right. \]  

(3.17)

To compute the element values for the lowpass prototype for a given passband ripple of \( L_{Ar} \) dB and with a cutoff frequency \( \Omega_c = 1 \), the following formulas may be used [31]:

\[ g_0 = 1 \]

\[ g_1 = \frac{2}{\gamma} \sin \left( \frac{\pi}{2n} \right) \]
3.2. CHEBYSHEV RESPONSE

\[ g_i = \frac{1}{g_{i-1}} \frac{4 \sin \left( \frac{(2i-1)\pi}{2n} \right)}{\gamma^2 + \sin^2 \left( \frac{(i-1)\pi}{n} \right)} \quad \text{for } i = 2, 3, \cdots, n \]

\[ g_{n+1} = \begin{cases} 
1.0 & \text{for } n \text{ odd} \\
\coth^2 \left( \frac{\beta}{4} \right) & \text{for } n \text{ even} 
\end{cases} \]

\[ \beta = \ln \left[ \coth \left( \frac{L_{A r}}{17.37} \right) \right] \]

\[ \gamma = \sinh \left( \frac{\beta}{2n} \right) \]

For a given passband ripple \( L_{A r} \) dB, a minimum stopband attenuation \( L_{A s} \) dB at \( \Omega = \Omega_s \), the degree of the lowpass prototype can be found by

\[ n \geq \cosh^{-1} \sqrt{\frac{(10^{0.1L_{A s}}) - 1}{(10^{0.1L_{A r}}) - 1}} \cosh^{-1} \Omega_s \]  \hspace{1cm} (3.18)

In this work, the filter response that is chosen for all filters is the Chebyshev response. The reason is that, as it will be explained later on, there exists the necessity to have a ripple behaviour in the passband. In some cases, it might be convenient to use the Butterworth response, that exhibits a flat behaviour in the passband. However, the Chebyshev response provides higher selectivity and simplicity in its realization. Moreover, the physical implementation of a Butterworth filter is unfeasible when it comes to the specifications required in this thesis, as it will be justified in section 6.2. To illustrate the differences between the Butterworth and the Chebyshev response, the lowpass prototypes of order \( n = 4 \) of both filters are shown in Figures 3.5 and 3.6, respectively. For the Chebyshev response, the ripple constant is \( \epsilon = 0.5 \) dB. The maximally flat response in the passband of the Butterworth response can be clearly seen. Nevertheless, with the Chebyshev filter, a faster cutoff rate is achieved.

Bild 3.5: Butterworth lowpass response.  \hspace{1cm} Bild 3.6: Chebyshev lowpass response.
4 Transmission line filters

In low-frequency filters, all elements explained in last chapter can be implemented with capacitors and inductors. However, when it comes to high frequency, the circuit theory is not valid anymore and other methods have to be brought up. Transmission line filters are those networks that approximate filters explained in last chapter.

From the high frequency point of view, a filter is a two-port network that intends to remove unwanted signal components while letting wanted ones through. Transmission line filters are microwave filters, which are designed to operate on signals in the Megahertz to Gigahertz frequency ranges which is the case of this work.

It is the objective of this chapter to discuss all filters that will appear in this text.

4.1 Immittance inverters

The basis of filter design have been presented in chapter 3. All this theory is valid for a very variety of filters. Nevertheless, it would be convenient to focus on the case that concerns this work, that is, high frequency electronic filters which are also known as microwave filters. In this section, a very important network to understand and analyze microwave filters is discussed.

4.1.1 Definitions

Immittance inverter is the general term to refer to either impedance or admittance inverters. An impedance inverter is a two-port network that has the property to invert the impedance connected in one port seen from the other port. That is, if one port is terminated with an impedance $Z_2$, the impedance $Z_1$ seen in the other port is

$$Z_1 = \frac{K^2}{Z_2}$$ (4.1)

where $K$ is real and known as the characteristic impedance of the inverter. As it can be seen in (4.1), if $Z_2$ is inductive, $Z_1$ will become conductive and the other way round.

Likewise, an admittance inverter is a two-port network where an admittance $Y_2$ connected on one port is seen, looking at the other port, as

$$Y_1 = \frac{J^2}{Y_2}$$ (4.2)

4.1.2 Filters with immittance inverters

It can be shown, as indicated in Figure 4.1, that a series inductance connected between two inverters is equivalent to a shunt capacitance seen from the exterior terminals. Simi-
larly, a series inductance is equivalent to a shunt capacitance in between two inverters, as shown in Figure 4.2.

Bild 4.1: Equivalence between a shunt capacitance and a series inductance with immittance inverters.

Bild 4.2: Equivalence between a series inductance and a shunt capacitance with immittance inverters.

Bild 4.3: Lowpass prototype filter with immittance inverters.

The lowpass prototypes structures in Figure 3.1 and in Figure 3.3 can be converted into the form shown in Figure 4.3 and in Figure 4.4, where the $g_i$ values are the prototype element values defined in section 3.1. The element values $Z_0$, $Z_{n+1}$, $L_{ai}$, $Y_0$, $Y_{n+1}$ and $C_{ai}$ are chosen arbitrarily and it will not affect on the filter response if $K_{i,i+1}$ and $J_{i,i+1}$ are set as indicated by equations (4.3) and (4.4) [31].

\[
K_{0,1} = \sqrt{\frac{Z_0 L_{a,1}}{g_0 g_1}} \quad K_{i,i+1} = \sqrt{\frac{L_{ai} L_{a(i+1)}}{g_i g_{i+1}}}_{i=1 \text{ to } n-1} \quad K_{n,n+1} = \sqrt{\frac{L_{an} Z_{n+1}}{g_n g_{n+1}}} 
\]

\[
J_{0,1} = \sqrt{\frac{Y_0 C_{a,1}}{g_0 g_1}} \quad J_{i,i+1} = \sqrt{\frac{C_{ai} C_{a(i+1)}}{g_i g_{i+1}}}_{i=1 \text{ to } n-1} \quad J_{n,n+1} = \sqrt{\frac{C_{an} Y_{n+1}}{g_n g_{n+1}}} 
\]
4.1. IMMITTANCE INVERTERS

As it has been explained before, the lowpass to bandpass mapping transforms the capacitance into a parallel $LC$ resonant circuit and the inductance into a series $LC$ resonant circuit. Which means that, to have a bandpass filter made of immittance inverters, the inductances in Figure 4.3 should be replaced by a series $LC$ circuit. Similarly, in Figure 4.4, parallel $LC$ circuits should be placed instead of shunt capacitances.

Afterwards, these lumped $LC$ resonant circuits have to be replaced by distributed circuits like, for example, microstrip resonators. In practice, the reactances or susceptances of the distributed circuits do not approximate those of the lumped resonators at all frequencies. Nevertheless, this is sufficient for some filters proposed in this work. Moreover, this non-ideal response will be exploited to implement certain shapes that will be explained in Chapter 6.

One of the simplest forms of inverters is a quarter-wavelength of transmission line. In (4.5) it is shown the impedance along the length of a transmission line with characteristic impedance $Z_0$ and loaded with an impedance $Z_L$.

$$
\frac{Z(l)}{Z_0} = \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)}
$$

By setting $l = \lambda/4$ in (4.5), the following expression is obtained

$$
\frac{Z_{in}}{Z_0} = \frac{Z_0}{Z_L}
$$

and solving the equation for $Z_{in}$,

$$
Z_{in} = \frac{Z_0^2}{Z_L}
$$

which means that a transmission line, with an impedance $Z_L$ connected in one side, exhibits an impedance $Z = Z_0^2/Z_L$ $\lambda/4$ away. Thus, considering one side of the transmission line to be port 1, the other side of the transmission line -which is at $\lambda/4$- to be port 2 and comparing (4.1) to (4.7) it is immediate to realize that a transmission line certainly acts like an impedance inverter with $Z_1 = Z_{in}$, $Z_2 = Z_L$ and $K = Z_0$.

It is now important to mention that this property is only valid for a certain frequency, since a transmission line is only $\lambda/4$ at some specific frequency. Despite this, quarter-wavelengths lines can be used satisfactorily as an immittance inverter in narrow-band filters. However, they do not work properly in broad-band filters.
4.2 Bandpass filters

4.2.1 Bandpass filters with open-circuited stubs

Bandpass filters can be designed with shunt short-circuited stubs that are $\lambda_{g0}/4$ long with connecting lines that are also $\lambda_{g0}/4$, where $\lambda_{g0}$ is the guided wavelength in the medium of propagation at the midband frequency $f_0$.

As the fields in microstrip lines propagate in two media -air and dielectric- it does not support a pure TEM (Transverse ElectroMagnetic) wave. Transverse modes occur in waveguides with homogeneous and isotropic materials and with metallic boundaries. Solving Maxwell’s equations [33] and taking into consideration the boundary conditions of the waveguide, the allowed modes can be found. Notwithstanding, the longitudinal components of a microstrip line may be neglected in front of the transverse components so the quasi-TEM approximation can be applied.

When the longitudinal components of the fields for the dominant mode remain very much smaller than the transverse components, they may be neglected and the quasi-TEM approximation can be applied.

In Figure 4.5, the appearance of a microstrip line is depicted. Where the conductor is on top of the dielectric and in open air and $t$, $L$ and $W$ account for the physical dimensions of the conductor: thickness, length and width, respectively. And $h$ denotes the thickness of the substrate whose dielectric constant is $\epsilon_r$.

Because of the inhomogeneous structure it is necessary to define the effective dielectric constant $\epsilon_{re}$, which is the first parameter that describes the transmission characteristics of microstrip lines. The characteristic impedance $Z_c$ is the other one. For very thin conductors, the expressions to obtain $\epsilon_{re}$ and $Z_c$, according to [31], are:

For $W/h \leq 1$:

$$
\epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left\{ \left( 1 + 12 \frac{h}{W} \right)^{-0.5} + 0.04 \left( 1 - \frac{W}{h} \right)^2 \right\}
$$

(4.8)
4.2. BANDPASS FILTERS

\[ Z_c = \frac{60}{\sqrt{\epsilon_{re}}} \ln \left( \frac{8h}{W} + 0.25 \frac{W}{h} \right) \]  
(4.9)

For \( W/h \geq 1 \):

\[ \epsilon_{re} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left( 1 + 12 \frac{h}{W} \right)^{-0.5} \]  
(4.10)

\[ Z_c = \frac{120\pi}{\sqrt{\epsilon_{re}}} \left\{ \frac{W}{h} + 1.393 + 0.677 \ln \left( \frac{W}{h} + 1.444 \right) \right\}^{-1} \]  
(4.11)

The guided wavelength can now be determined and it is given by

\[ \lambda_{g0} = \frac{\lambda_0}{\sqrt{\epsilon_{re}}} \]  
(4.12)

where \( \lambda_0 \) is the free space wavelength at the operation frequency \( f_0 \). It can also be expressed as follows:

\[ \lambda_g = \frac{c}{f_0 \sqrt{\epsilon_{re}}} \]  
(4.13)

where \( c \) is the speed of light in the vacuum of free space and is \( c \approx 3 \cdot 10^8 \) m/s.

Now, the physical lengths of the stubs and the connecting lines can be determined. These kind of bandpass filters, made of shunt short-circuited stubs that are \( \lambda_{g0}/4 \) long with connecting lines that are also \( \lambda_{g0}/4 \) can be explained by looking at the impedance expression [33]:

\[ \frac{Z(l)}{Z_0} = \frac{Z_L + jZ_0 \tan(\beta l)}{Z_0 + jZ_L \tan(\beta l)} \]  
(4.14)

where \( Z(l) \) is the impedance along the transmission line, \( Z_0 \) is its characteristics impedance, \( Z_L \) is the load impedance and \( \beta \) is the propagation constant, which can be determined by:

\[ \beta = \frac{2\pi}{\lambda_g} \]  
(4.15)

As the short-circuited (\( Z_L = 0 \)) stubs are \( \lambda_{g0}/4 \) long, \( l \) has to be set to \( l = \lambda_{g0}/4 \). Then, the product \( \beta l \) equals to \( \pi/2 \) which makes the impedance, seen from the connecting lines, infinity. That is, the stubs look like open circuits so they have no influence on the structure. Thus, the midband frequency \( f_0 \) is not attenuated. At frequencies far from resonance, the impedance of the filter components change, and the signal is attenuated.

The quarter-wavelength connecting lines implement the immittance inverters explained before, while the quarter-wavelength short-circuited stubs approximate the reactances of the \( LC \) resonant circuits in near resonance. This is only an approximation, nevertheless, it is sufficient to implement the bandpass filters, needed in this work, for the frequency range from 6 GHz to 8.5 GHz. The series \( LC \) circuits connected between impedance inverters are replaced by the reactance of the distributed resonator. Similarly, the shunt \( LC \) circuits connected between admittance inverters are replaced by the susceptance of the distributed resonator. This can be done by making the distributed resonator reactance/susceptance and the reactance/susceptance slope (frequency
derivative of the reactance/susceptance) equal to their corresponding lumped-resonator values at band center.

Nonetheless, short-circuited stubs are difficult to implement since via-hole groundings are needed. All microstrip stubs have to be pierced and soldered to the ground plane, which may lead to imprecisions, errors and degradation of the substrate and, hence, a defective filter response.

This soldering, and all its consequences, can be avoided by replacing the \( \lambda_{g0}/4 \) short-circuited stubs by \( \lambda_{g0}/2 \) open-circuited stubs. According to (4.14), an open-circuited \( Z_L = \infty \) stub whose length is \( l = \lambda_{g0}/2 \) exhibits the same impedance, near resonance, in the connecting lines, that is, the resultant filter will have similar passband characteristics as the one with \( \lambda_{g0}/4 \) short-circuited stubs. Nevertheless, the stopband will have attenuation poles at the frequencies \( f_0/2 \) and \( 3f_0/2 \) and it will have additional passbands in the vicinity of \( f = 0 \) and \( f = 2f_0 \), and at other corresponding periodic frequencies. In forthcoming chapters, this issue will be taken up again.

It can be shown that short stubs can approximate shunt capacitors. According to (4.14), the input admittance of an open-circuited stub of lossless microstrip line having characteristic admittance \( Y_c = 1/Z_c \) and propagation constant \( \beta = 2\pi/\lambda_g \) is

\[
Y_{in} = jY_c \tan \left( \frac{2\pi l}{\lambda_g} \right) \tag{4.16}
\]

where \( l \) is the length of the stub. It can be noted that the input admittance is capacitive. If the stub is shorter, \( l < \lambda_g/8 \), the value of \( \tan(2\pi l/\lambda_g) \) may be approximated by its argument \( 2\pi l/\lambda_g \). This way, the admittance results in

\[
Y_{in} \approx jY_c \left( \frac{2\pi l}{\lambda_g} \right) = j\omega \left( \frac{Y_c l}{v_p} \right) \tag{4.17}
\]

So, a short open-circuited stub is equivalent to a shunt capacitance \(^1\) \( C = Y_c l/v_p \) where \( v_p \) is the phase velocity of propagation in the stub.

For a given filter degree \( n \), the bandpass filter characteristics will depend on the characteristic impedances of the stub lines \( (Z_i) \) and on the characteristic impedances of the connecting lines \( (Z_{i,i+1}) \). The design equations to determine all these characteristic impedances are given by [31]

\[
\begin{align*}
\omega_0 &= \frac{\omega_1 + \omega_2}{2} & \omega &= \frac{\omega_2 - \omega_1}{\omega_0} \\
\theta &= \frac{\pi}{2} \left( 1 - \frac{\omega}{2} \right) \\
J_{1,2} &= \frac{g_0}{Z_0} \sqrt{\frac{hg_1}{g_2}} & J_{n-1,n} &= \frac{g_0}{Z_0} \sqrt{\frac{hg_1 g_{n+1}}{g_0 g_{n-1}}} \\
J_{i,i+1} &= \frac{hg_0 g_{i+1}}{Z_0 \sqrt{g_i g_{i+1}}} \quad \text{for } i = 2 \text{ to } n - 2
\end{align*}
\]

\(^1\)The admittance of a capacitor with capacitance \( C \) farads is given by \( Y = j\omega C \) siemens
4.2. BANDPASS FILTERS

\[ N_{i,i+1} = \sqrt{(Z_0 J_{i,i+1})^2 + \left(\frac{h g_1 \tan(\theta)}{2}\right)^2} \quad \text{for } i = 1 \text{ to } n - 1 \]

\[ Z_1' = \left\{ \frac{g_0}{Z_0} \left(1 - \frac{h}{2}\right) g_1 \tan \theta + \frac{1}{Z_0} (N_{1,2} - Z_0 J_{1,2}) \right\}^{-1} \]

\[ Z_n' = \left\{ \frac{1}{Z_0} \left(g_n g_{n+1} - g_0 g_1 \frac{h}{2}\right) \tan \theta + \frac{1}{Z_0} (N_{n-1,n} - Z_0 J_{n-1,n}) \right\}^{-1} \]

\[ Z_i' = \left\{ \frac{1}{Z_0} (N_{i-1,i} + N_{i,i+1} - Z_0 J_{i-1,i} - Z_0 J_{i,i+1}) \right\}^{-1} \quad \text{for } i = 2 \text{ to } n - 1 \]

\[ Z_i = \frac{2Z_i' \tan^2(\theta) - 1}{\tan^2(\theta) - 1} \quad \text{for } i = 1 \text{ to } n \]

\[ Z_{i,i+1} = \frac{1}{J_{i,i+1}} \quad \text{for } i = 1 \text{ to } n - 1 \]

where \( \omega_1 \) and \( \omega_2 \) are the lower and higher angular cutoff frequencies, \( g_i \) are the element values of the lowpass prototype filter given for a normalized cutoff \( \Omega_c = 1 \), \( h \) is a dimensionless constant (typically chosen to be 2) which can be chosen so as to give a convenient impedance level in the interior of the filter. All stubs are \( \lambda g_0/2 \) and all connecting lines are \( \lambda g_0/4 \), where \( \lambda g_0 \) is the wavelength in the medium of propagation at the mid-band frequency \( f_0 \), which can be computed with (4.8), (4.10) and (4.13).

For this type of filters, the impedance of the end stubs are much greater than the impedance of the stubs in the interior of the filter. For this reason it is sometimes convenient to build this type of filter with double stubs in the interior of the filter and with single stubs at the ends [28].

4.2.2 Bandpass filter based on the optimum distributed highpass filter

In the previous section, a bandpass filter using open-circuited stubs has been discussed. Due to the fact that these filters are made out of impedance inverters, they only behave as inverters in a certain frequency range which is not sufficient for our purposes since our band of interest is as wide as 7500 MHz. This presents a new challenge to the bandpass filter design, not only for this huge bandwidth, a Ultra-Wideband bandpass filter should exhibit very good selectivity in order to meet the FCC spectrum mask. Therefore, a high-order filter design is required, although it is also needed to guarantee low insertion loss and not too large circuit size. In addition, flat group delay response is wanted to have minimum distortion to the input pulse. To fulfill this requirements, the optimum distributed highpass filter structure [29] is adopted for bandpass filter design.

We are going to take advantage of the periodicity in the filtering characteristics of the highpass filter in such a way that the second attenuation pole due to periodicity falls in the upper cutoff frequency of the bandpass that we would need.

This type of filter consists of a cascade of shunt short-circuited stubs of electrical length \( \theta_c \) at the lower cutoff frequency \( f_c \) of the passband, separated by connecting lines
of electrical length $2\theta_c$. With this filter, a size reduction is achieved while preserving the properties of larger filters. This is because, although the filter consists of only $n$ stubs, it has an insertion function of degree $2n - 1$ in frequency. Therefore, this filter will have a faster rate of cutoff than those bandpass filters of order $n$ with $n$ stubs discussed in 4.2.1. For bandpass applications, the filter has a primary passband from $\theta_c$ to $\pi - \theta_c$ with a cutoff at $\theta_c$. The harmonic passbands occur periodically, centered at $\theta = 3\pi/2, 5\pi/2, \ldots$, and separated by attenuation poles located at $\theta = \pi, 2\pi, \ldots$. In Table 4.1, some typical element values of the network in Figure 4.6 are tabulated. These values are for the practical design of the filter using from two to six stubs and a passband ripple of 0.1 dB for $\theta_c = 25^\circ, 30^\circ$, and $35^\circ$. The tabulated elements are the normalized characteristic admittances of transmission line elements. Therefore, given the terminating impedance $Z_0$, the following operation should be applied to obtain the actual line impedances:

$$Z_i = Z_0/y_i$$
$$Z_{i,i+1} = Z_0/y_{i,i+1}$$

(4.18)

Interpolation may be done in order to obtain the normalized admittances $y_i$ and $y_{i,i+1}$ for values of $\theta_c$ different from $25^\circ, 30^\circ$ and $35^\circ$. An example of this method will be shown later.
4.2. BANDPASS FILTERS

4.2.3 Parallel-coupled bandpass filter

The third and last kind of bandpass filter used in this work is the parallel-coupled microstrip bandpass filter. As it has been explained before, a bandpass filter can be implemented using immittance inverter with series LC resonant circuits. Which corresponds to Figure 4.1 and applying the frequency transformation that replaces the series inductances by series LC resonant circuits.

One possible method to implement this cascade of immittance inverter and series LC circuit is by means of parallel-coupled lines. Figure 4.8 illustrates the equivalence between coupled transmission lines and the required circuit structure.

<table>
<thead>
<tr>
<th>n</th>
<th>$\theta_c$</th>
<th>$y_1$</th>
<th>$y_{1,2}$</th>
<th>$y_2$</th>
<th>$y_{2,3}$</th>
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<td>$25^\circ$</td>
<td>0.15436</td>
<td>1.13482</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$30^\circ$</td>
<td>0.22070</td>
<td>1.11597</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$35^\circ$</td>
<td>0.30755</td>
<td>1.08967</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>$25^\circ$</td>
<td>0.19690</td>
<td>1.12075</td>
<td>0.18176</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$30^\circ$</td>
<td>0.28620</td>
<td>1.09220</td>
<td>0.30726</td>
<td></td>
<td></td>
</tr>
<tr>
<td></td>
<td>$35^\circ$</td>
<td>0.40104</td>
<td>1.05378</td>
<td>0.48294</td>
<td></td>
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</tr>
<tr>
<td>4</td>
<td>$25^\circ$</td>
<td>0.22441</td>
<td>1.11113</td>
<td>0.23732</td>
<td>1.10361</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$30^\circ$</td>
<td>0.32300</td>
<td>1.07842</td>
<td>0.39443</td>
<td>1.06888</td>
<td></td>
</tr>
<tr>
<td></td>
<td>$35^\circ$</td>
<td>0.44670</td>
<td>1.03622</td>
<td>0.60527</td>
<td>1.01536</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>$25^\circ$</td>
<td>0.24068</td>
<td>1.10540</td>
<td>0.27110</td>
<td>1.09317</td>
<td>0.29659</td>
</tr>
<tr>
<td></td>
<td>$30^\circ$</td>
<td>0.34252</td>
<td>1.07119</td>
<td>0.43985</td>
<td>1.05095</td>
<td>0.48284</td>
</tr>
<tr>
<td></td>
<td>$35^\circ$</td>
<td>0.46895</td>
<td>1.02790</td>
<td>0.66089</td>
<td>0.99884</td>
<td>0.72424</td>
</tr>
<tr>
<td>6</td>
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<td>0.33031</td>
</tr>
<tr>
<td></td>
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<td>0.35346</td>
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<tr>
<td></td>
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<td>1.02354</td>
<td>0.68833</td>
<td>0.99126</td>
<td>0.77546</td>
</tr>
</tbody>
</table>

Tabelle 4.1: Element values of optimum distributed highpass filters with 0.1 dB ripple. Source [31].

Bild 4.8: Equivalent circuit of two coupled transmission lines.
4.2. BANDPASS FILTERS

The reactance $X_{eq}$ and the impedance inverter parameter $K$ referred in Figure 4.8 are given by [30]

$$X_{eq} = \frac{\pi Z_e + Z_o}{4} \tag{4.19}$$

$$K = \frac{Z_e - Z_o}{2} \tag{4.20}$$

where $Z_e$ and $Z_o$ are the characteristic impedance of the even and of the odd mode, respectively. The coupled line structure supports two quasi-TEM modes, the even mode and the odd mode. For an even-mode excitation, both microstrip lines have the same voltage potentials while in the odd mode, both lines have the opposite voltage potentials. The even mode can be modelled with a magnetic wall at the symmetry plane and in the odd mode, with an electric wall. In general, these two modes will be excited at the same time. However, as they propagate with different phase velocities, they experience different permittivities. Therefore, the coupled lines are characterized by the characteristic impedances as well as the effective dielectric constants for the two modes.

With all this, the filter structure will consist of half-wavelength resonators that are positioned so that adjacent resonators are parallel to each other along half of their length as shown in Figure 4.9.

![Bild 4.9: General structure of parallel-coupled microstrip bandpass filter](image)

The dual configuration can be considered as well. Instead of series $LC$ resonant circuits in between impedance inverters, shunt $LC$ resonators in between admittance inverters
4.3. BANDSTOP FILTERS

also have the same filter response. The design equations for this coupled lines filter are given by [31]

\[ \frac{J_{01}}{Y_0} = \sqrt{\frac{\pi FBW}{2 g_0 g_1}} \] (4.21)

\[ \frac{J_{j,j+1}}{Y_0} = \frac{\pi FBW}{2} \frac{1}{g_j g_{j+1}} \quad j = 1 \text{ to } n - 1 \] (4.22)

\[ \frac{J_{n,n+1}}{Y_0} = \sqrt{\frac{\pi FBW}{2 g_n g_{n+1}}} \] (4.23)

where \( g_0, g_1, \ldots, g_n \) are the elements of the lowpass prototype with a normalized cutoff \( \Omega_c = 1 \), and \( FBW \) is the fractional bandwidth of bandpass filter. \( J_{j,j+1} \) are the characteristic admittances of \( J \)-inverters and \( Y_0 \) is the characteristic admittance of the terminating lines, that is, \( Y_0 = 1/Z_0 \). The realization of the \( J \)-inverters is obtained through the even- and odd-mode characteristic impedances of the coupled microstrip line resonators, which can be computed with [31]

\[ (Z_{0e})_{j,j+1} = \frac{1}{Y_0} \left[ 1 + \frac{J_{j,j+1}}{Y_0} + \left( \frac{J_{j,j+1}}{Y_0} \right)^2 \right] \quad j = 0 \text{ to } n \] (4.24)

\[ (Z_{0o})_{j,j+1} = \frac{1}{Y_0} \left[ 1 + \frac{J_{j,j+1}}{Y_0} - \left( \frac{J_{j,j+1}}{Y_0} \right)^2 \right] \quad j = 0 \text{ to } n \] (4.25)

Once \( (Z_{0e})_{j,j+1} \) and \( (Z_{0o})_{j,j+1} \) have been calculated, the widths \( W_j \) and the spacing between lines \( s_j \) that exhibit such even- and odd-mode impedances, have to be found. These values can be computed using equations in [31].

The coupled lines are implemented using microstrip technology. The sketch of microstrip coupled lines is shown in Figure 4.10. Where \( t \) accounts for the thickness of the metal that compounds the ground plane and the conductor, \( W \) is the line width, \( L \) the length of the line, \( h \) the thickness of the dielectric with constant \( \epsilon_r \) and \( s \) denotes the spacing between lines.

4.3 Bandstop filters

Bandstop filters play the main role of filtering out the unwanted signals and passing the desired signals. In this section, two types of bandstop filters will be presented. First, conventional bandstop filters which have a key disadvantage that prevent us from using them for Ultra-Wideband applications. Second, bandstop filters with extended upper passbands that solve this problem. Both filters with their advantages and disadvantages are explained in detail in the next two subsections.
4.3. BANDSTOP FILTERS

4.3.1 Conventional bandstop filters

Conventional bandstop filters consist of quarter-wavelength open-circuited stubs separated by connecting lines that are a quarter wavelength long at the mid-stopband frequency. As the lengths are fully defined, the following step is to find out the characteristic impedances of the connecting lines and the characteristic impedances of the stubs that approximate the behaviour of the lumped elements that conform a bandstop filter. Nevertheless, this method uses unit elements that are redundant, and their filtering properties are not fully utilized, so that in this sense, the filter is not an optimum one. To design an optimum bandstop filter, tabulated element values are available in [31] for filters with \( n \) stubs, with a certain passband ripple \( \epsilon \) dB, a desired fractional bandwidth \( FBW \) and a passband return loss of \(-20\) dB. This optimum filter achieves steeper attenuation and more uniform stub line widths. Moreover, a redundant filter may have extremely narrow outer stubs that can become unrealizable.

However, this filters have a severe restriction on the extent of the upper passband imposed by the periodicity of the distributed elements. They present spurious stop bands periodically centered at frequencies that are odd multiples of the midband frequency \( f_0 \). At these frequencies, the shunt open-circuited stubs in the filter are odd multiples of \( \lambda_{g0}/4 \) long, with \( \lambda_{g0} \) being the guided wavelength at frequency \( f_0 \) so that they short out the main line and cause spurious bands. The first upper stopband center frequency is located at three times the fundamental stopband.

Particularizing to the case of this thesis, the bandstop filter will be used to remove two bandwidths with midband frequencies \( f_0 = 2.5 \) GHz and \( f_0 = 3.5 \) GHz. This band removal will be performed for the EC mask case, therefore, the band comprised from 6 GHz to 8.5 GHz must not be attenuated. It has to be noted that the first spurious band occur at \( 3f_0 = 7.5 \) GHz. That is exactly the band of interest. This fact makes the use of this type of bandstop filter unfeasible and requires the introduction of the following section.
4.3.2 Extended upper-passband bandstop filters

This new design [32] remove the repeated odd-multiple resonance limitation using non-uniform stepped impedance resonators that have shorter electrical lengths. This passband extension is accomplished by replacing each branch with quarter-wave length (\(\lambda/4\)) with its equivalent two-section stepped impedance shaped section as illustrated in Figure 4.11. Denoting with \(\theta_1\) and \(\theta_2\) the effective electrical lengths of the lines with \(Z_1\) and \(Z_2\) characteristics impedances, respectively, the following equation can be found:

\[
K = \frac{Z_2}{Z_1} = \tan(\theta_1) \tan(\theta_2)
\]  

(4.26)

The ratio of the first spurious response to the center frequency \(f_{s1}/f_0\) depends on the values of the impedance ratio \(K\) as well as the stepped-sections electrical lengths.

![Bild 4.11: Equivalence of stepped impedance and open circuit sections.](image)

![Bild 4.12: Variation of the first spurious frequency with the impedance ratio \(K\). Source: [32].](image)

The design procedure is as follows. First of all, the midband frequency \(f_0\) of the stopband has to be chosen. Afterwards, it is needed to know how far the first spurious response \(f_{s1}\) should be allocated. Computing \(f_{s1}/f_0\) and the ratio between the two characteristic impedances of the stepped section \(K = Z_2/Z_1\), the first effective electrical length \(\theta_1\) can be obtained by looking at Figure 4.12. With \(\theta_1\) and \(K\) it is possible to find out the value of \(\theta_2\) in Figure 4.13.
### 4.4 Lowpass filters

It is also required for the goal of this work to implement a lowpass filter. The design of a lowpass filter starts by selecting an appropriate lowpass prototype, such as one as described in section 3.1. According to the specifications some parameters will have to be chosen: the passband ripple $\epsilon$, the number of active elements or the degree of the filter $n$, and the type of response (Butterworth, Chebyshev, ...). Then, the element values of the lowpass prototype, which are normalized to make a source impedance $g_0 = 1$ and a cutoff frequency $\Omega_c = 1$, are obtained. Next, they have to be transformed to the L-C elements for the desired cutoff frequency $\omega_c$ and the desired source impedance, which is normally $Z_0 = 50$ ohms for microstrip filters. The next step is to find a microstrip realization that approximates the lumped elements. In this concrete case, the L-C elements will be approximated using the stepped-impedance technique.

#### 4.4.1 Stepped-impedance lowpass filters

The stepped-impedance technique tries to approximate lumped elements by using a cascaded structure of alternating high- and low-impedance transmission line. The general structure is shown in Figure 4.14. These segments have to be much shorter than the wavelength in order to be act as semilumped elements and approximate the circuit in Figure 4.15.

![Bild 4.14: General structure of the stepped-impedance lowpass microstrip filters](image)

Bild 4.13: $\theta_1$ versus $\theta_2$ for different values of $K$. Source: [32].
So that the transmission lines act like semilumped elements, the characteristic impedance of the low impedance line should be lower than the source impedance, which is usually 50 ohms for microstrip lines. And the high impedance line should be greater than the source impedance line. That is $Z_{0C} < Z_0 < Z_{0L}$, where $Z_{0C}$ denotes the characteristic impedance of the low impedance line that acts like a shunt capacitor and $Z_{0L}$ is the characteristic impedance of the high impedance line that acts like a series inductor.

To see how the transmission lines approximate the lumped elements, the Z-parameters should be introduced. A two-port network can be represented as illustrated in Figure 4.16 and in equations (4.27).

$$\begin{align*}
V_1 &= Z_{11}I_1 + Z_{12}I_2 \\
V_2 &= Z_{21}I_1 + Z_{22}I_2
\end{align*}$$

(4.27)

Z-parameters for transmission lines having characteristic impedance $Z_0$ are given by [33]

$$\begin{align*}
Z_{11} &= Z_{22} = -jZ_0 \cot(\beta l) \\
Z_{12} &= Z_{21} = -jZ_0 \csc(\beta l)
\end{align*}$$

(4.28)  
(4.29)
Then, the series element \( Z_{11} - Z_{12} (= Z_{22} - Z_{21}) \) are given by

\[
Z_{11} - Z_{12} = -jZ_0 \frac{\cos(\beta l)}{\sin(\beta l)} + jZ_0 \frac{1}{\sin(\beta l)} \tag{4.30}
\]

\[
= jZ_0 \frac{1 - \cos(\beta l)}{\sin(\beta l)} \tag{4.31}
\]

\[
= -jZ_0 \frac{\cos(\beta l) - 1}{\sin(\beta l)} \tag{4.32}
\]

\[
= jZ_0 \tan\left(\frac{\beta l}{2}\right) \tag{4.33}
\]

and the shunt element is simply \( Z_{12} (= Z_{21}) \). The element values for the Figure 4.17 are:

\[
\frac{X}{2} = Z_0 \tan\left(\frac{\beta l}{2}\right) \tag{4.34}
\]

\[
B = \frac{\sin(\beta l)}{Z_0} \tag{4.35}
\]

Now, assuming that length \( l \) is much shorter than wavelength \( (l < \lambda/8) \) and that the characteristic impedance is large, the sine can be approximated by its argument and the cosine by one, then

\[
X \approx Z_0 \beta l \tag{4.36}
\]

\[
B \approx 0 \tag{4.37}
\]
4.4. LOWPASS FILTERS

\[ jX \quad jB \]

\[ \frac{B}{jX} \]

Bild 4.18: Equivalent circuit for large and small Zo respectively.

Which implies that it is a series inductor. Similarly, assuming very small characteristic impedance it results in a shunt capacitor.

\[ X \approx 0 \quad (4.38) \]
\[ B \approx Y_0/\beta l \quad (4.39) \]

This assumptions are summarized in Figure 4.18.

A lower \( Z_0 \) results in a better approximation of a shunt capacitor. Nevertheless, the resulting line width must not allow any transverse resonance to occur. On the other hand, a higher \( Z_0 \) means a better approximation of a series inductor, but a high impedance leads to a very narrow line which might be difficult to fabricate.

The impedances \( Z_{0C} \) and \( Z_{0L} \) (or the lines widths \( W_C \) and \( W_L \)) have to be set taking into account the above-mentioned considerations. Then, the physical lengths of the high- and low-impedance lines may be found by [31]

\[ l_L = \frac{\lambda_{sl}}{2\pi} \sin^{-1} \left( \frac{\omega c}{Z_{0L}} \right) \]
\[ l_C = \frac{\lambda_{sc}}{2\pi} \sin^{-1} \left( \frac{\omega c Z_{0C}}{Z_{0C}} \right) \quad (4.40) \]
5 Basic concepts for filter analysis

In this chapter, several ways to evaluate the behaviour of the filter are presented. This work concentrates and is mainly based on methods that are designed and presented in the frequency domain. Even so, it is also important to have a look at the time domain since it is necessary to know how the output waveform looks like in the time domain and how the filter affects the whole system performance and the signal itself.

5.1 Frequency domain analysis

The mathematical expression of $S_{21}$ parameter explained in chapter 3 corresponds to the filter transfer function. This term (transfer function) is often used exclusively to refer to linear, time-invariant systems such as filters. In the frequency domain, the signal is called spectrum and it is obtained computing the Fourier transform of the time domain waveform. The Fourier transform may be interpreted as the decomposition of a signal in frequency components. That is, any waveform can be expressed as a sum of sinusoids and the Fourier transform describes how much of the signal lies within each given sinusoid.

The scenario under consideration is drawn in Figure 5.1. Let $X(f)$ be an input signal passing through a linear, time-invariant (LTI) system $H(f)$ and $Y(f)$ the resulting output. The transfer function $H(f)$ relates the input spectrum and the output spectrum as

$$Y(f) = H(f)X(f)$$

(5.1)

The most common used parameters in the frequency domain [31] can be obtained from the transfer function. The insertion loss response of the filter can by computed by

$$L_A(f) = 10 \log \frac{1}{|H(f)|^2} [dB]$$

(5.2)

which is the loss of signal power resulting from the insertion of a filter in a transmission line. Furthermore, as $|S_{11}|^2 + |S_{21}|^2 = 1$ for a lossless, passive two-port network, the return loss response of the filter is

$$L_R(f) = 10 \log[1 - |H(f)|^2] [dB]$$

(5.3)
where $S_{11}$ corresponds to the reflexion coefficient, $S_{21}$ is the transfer function $H(f)$ and $L_R(f)$ is the reflection of signal power resulting from the insertion of the filter in a transmission line.

The transfer function may be rewritten as shown in (5.4)

$$H(f) = |H(f)| e^{j\phi(f)}$$

(5.4)

where $\phi(f) = \arg(H(f))$ is the phase of $H(f)$ and $|H(f)|$ is called magnitude.

The group delay response of the filter can be calculated by

$$\tau_d(f) = -\frac{1}{2\pi} \frac{d\phi(f)}{df}$$

(5.5)

It is desirable for the group delay to be constant all over the frequency range; otherwise the signal is distorted. As defined in (5.5), a constant group delay can be achieved if the transfer function has a linear phase response. The degree of nonlinearity of the phase indicates the deviation of the group delay from a constant. The group delay is expressed in units of time and gives an indication of the time delay that the impulse signal suffers. In other words, it is a measure of how long it takes a signal to traverse the network. The degree of the filter has big impact on the group delay: the higher the filter order, the more group delay a filter introduces.

### 5.2 Time domain analysis

Similarly as in the frequency domain, the scenario is comprised of the input waveform $x(t)$ that is applied at the filter which responds with the output waveform $y(t)$. This situation is described in Figure 5.2

![Bild 5.2: LTI system in the time domain.](image)

In order to compute the output $y(t)$, let us assume that a linear, time-invariant system has no energy stored and an excitation $x(t) = \delta(t)$ is applied, that is, the unit impulse function, at $t = 0$. The response of the system for this case is known as impulse response denoted with $h(t)$. As it is a LTI system, the response to $a\delta(t - t_1) + b\delta(t - t_2)$ will be $ah(t - t_1) + bh(t - t_2)$. An arbitrary signal can be seen as many unit impulse functions multiplied by different values each. If these unit impulse functions are denoted by $\delta(t - \tau)$, where $\tau$ is the variable that moves the impulse functions along the time axis, and $x(\tau)$ is the amplitude for each $\tau$, and taking into account that the system is linear and time-invariant, the output is the sum of all outputs for each excitation. Because the duration of each impulse tends to zero, the sum is a integration and the excitation is

$$x(t) = \int_{-\infty}^{\infty} x(\tau)\delta(t - \tau)d\tau$$

(5.6)
and the output
\[ y(t) = \int_{-\infty}^{\infty} x(\tau) h(t - \tau) d\tau \] (5.7)

This integration (5.7) is called convolution and it is denoted with the operator \(*\).

In short, the output \( y(t) \) is obtained with the convolution between the input \( x(t) \) and the filter impulse response \( h(t) \) as expressed in (5.8).

\[ y(t) = x(t) * h(t) = \int_{-\infty}^{\infty} x(\tau) h(t - \tau) d\tau = \int_{-\infty}^{\infty} x(t - \tau) h(\tau) d\tau \] (5.8)

When the available data are not the measurements in time domain but in frequency domain, as it will be the case, the strategy is to obtain them indirectly by means of the inverse Fourier transform [18].

First of all, the analytic function, in the frequency domain, has to be defined
\[ H(f) = \begin{cases} 2H(f) & \text{for } f > 0 \\ 0 & \text{for } f \leq 0 \end{cases} \] (5.9)

where \( H(f) \) is the transfer function from Figure 5.1.

The inverse Fourier transform of (5.9) gives the complex analytical response \( h^+(t) \) in the time domain. The impulse response \( h(t) \) of the filter is the real part of the complex analytical response \( h^+(t) \), that is
\[ h(t) = \Re \{ h^+(t) \} \] (5.10)

Common time-domain measures [19] are the pulse width and ringing of the envelope of the analytic response \( |h^+(t)| \). The first parameter, that has to be defined to compute the rest, is the peak value \( P \), defined as
\[ P = \max_t |h^+(t)| \] (5.11)

Then, to measure the linear distortion that the filter introduces, the envelope width is used. It is defined as the full width at half maximum (FWHM) of the magnitude of the impulse response envelope:
\[ w_{0.5} = t_2 \|h^+(t_2)|=P/2 - t_1 \|t_1<t_2,|h^+(t_1)|=P/2 \] (5.12)

The second measure is the duration of the ringing \( \tau_{r,\alpha} \), that is defined as the time until the envelope of the impulse response has fallen from the peak value \( P \) below a certain lower bound. That is, below a fraction \( \alpha \) of \( P \)
\[ \tau_{r,\alpha} = t_\alpha \|h^+(t_\alpha)|=\alpha P - t_P \|t_P>t_\alpha,|h^+(t_P)|=P \] (5.13)
In this work, the lower bound will be set to 10% of the peak value $P$. That is, $\alpha = 0.1$. In Figure 5.3, the maximum peak value, the envelope width (FWHM) and the duration of the ringing are shown.
6 Filter design for the Vivaldi antenna

The filter design will be addressed in this chapter. All filters will be planar and realized in microstrip technique. And, because the strip conductor is exposed on the top side, component mounting is relatively easy. The dielectric constant $\varepsilon_r$ and the thickness $h$ of the substrate (see Figure 4.5) will be chosen depending on the characteristics of the filters as it will be explained in the following sections. The thickness $t$ of the conductor will be 17 $\mu$m in all cases.

In this chapter, the antenna under study is the Vivaldi antenna whose transfer function is plotted in Figures 6.1 and 6.2. The blue curve corresponds to the measured transfer function of the antenna in the main beam direction, co-polarized component, while the red curve is the smoothed version of the blue curve. As it can be seen, the measured transfer function can be approximated with a straight line with decreasing slope. The linear approximation is used to compute the expression (2.3), and to obtain the $S_{21}$ parameter that the pre-distortion filter should have, without taking into account the rough variations that cannot be compensated. In Figure 6.1, the antenna transfer function is plotted for the FCC frequency range, while in Figure 6.2 the transfer function is plotted for the EC frequency range.
6.1 Filter for the FCC mask and the Vivaldi antenna

6.1.1 Shaping filter

The aim of this section is to find a microstrip filter whose transfer function, denoted as \( H(f) \), is as close as possible to that filter response depicted in Figure 6.3. Taking into account that the antenna transfer function \( A(f) \) is the one plotted in red in Figure 2.8 and that the desired antenna output must be flat between 3.1 GHz to 10.6 GHz, as the FCC mask demands (see Figure 1.1).

Two different ways to approximate the required shape could be used. First, designing a filter whose passband has a single ripple which should be as deep as the target response. Unfortunately, this option has to be discarded. The required ripple is of the order of 15 dB (Figure 2.13) and it should be as wide as 7.5 GHz which is completely unfeasible. The second -and adopted- method is to take advantage of the transition band\(^1\) of the bandpass filter. The main issue is the following: a high attenuation level (around 30 dB) is needed at \( f = 3.1 \) GHz while 0 dB at \( f = 10.6 \) GHz. Which means that the transition band should comprise a very wide bandwidth. This cannot be achieved with conventional bandpass filters since the transition band of the lowest order is much steeper than the required filter response.

The most suitable microstrip structure for this purpose is a parallel-coupled microstrip filter. This kind of filters are intended for implementing bandpass filters. They originally make use of half-wavelength resonators that are positioned so that adjacent resonators are parallel to each other along half of their length. This filter is discussed in section 4.2.3 where the design equations are also provided. Nevertheless, this structure is going to be modified and adapted to the case that concerns us.

Thereby, a modified version of a parallel-coupled filter has to be used. The spacing between lines is increased and their lengths modified by the CST optimizer in order

\(^1\)Transition between the stopband and the passband
to have a smoother transition band that can approximate satisfactorily the desired $S_{12}$ parameter. The number of coupled lines is the degree of the filter. Therefore, the more resonant coupled lines, the steeper the filter response is. Hence, in order to obtain the desired attenuation, after experimenting with simulations, it has been found out that the required filter order is 2. Consequently, two coupled lines are needed.

![CST layout of the shaping filter (left) and its simulated $S_{21}$ parameters.](image)

**Bild 6.4:** CST layout of the shaping filter (left) and its simulated $S_{21}$ parameters.

The microstrip filter is then simulated on a substrate with a relative dielectric constant of 2.2 whose thickness is 1.57 mm and a metal thickness of 17 $\mu$m. The layout of the filter and the electromagnetic (EM) simulated performance of the filter compared to the ideal filter response is shown in Figure 6.4. A good resemblance to the wanted shape is obtained within the band of interest, from 3.1 GHz to 10.6 GHz. The next step is to simulate the performance of the whole system in Figure 2.5 using the proposed structure in Figure 6.4 as the filter $H(f)$. In Figure 6.5 is shown the computed on-the-air emitted spectrum $S(f)$ when the filter is used, where $S(f)$ is the spectrum introduced in Chapter 2.

As expected, the whole system exhibits a good behaviour within the band. For comparison, both the antenna output $S(f)$ with the proposed filter and the output without the filter is provided in Figure 6.5. The band occupancy has been increased considerably with the parallel-coupled structure. However, the spectrum still violates the mask below 3.1 GHz and beyond 10.6 GHz. This means that a bandpass filter is needed in order to select the desired UWB band. The addition of a bandpass filter can select the band of interest by attenuating the frequencies outside it. Nevertheless, the mask violation is more severe at the low frequencies, where the output is above the mask only 1.18 dB. So the use of a bandstop filter should be also considered to filter out the frequencies from 2 GHz to 3.1 GHz. So, two different methods have been studied and measured in the following sections.
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

6.1.2 Shaping filter with bandpass filter

The first option that is analyzed is the introduction of a bandpass filter. The design procedure is as follows. From simulations, it has been derived that six short-circuited stubs provide sufficient selectivity for filtering out the unwanted signal. The implemented bandpass filter is based on the optimum distributed highpass filter. Although it only consists of six stubs, the order of the filter is \( n = 11 \) which will provide sufficient selectivity. Once the two cutoff frequencies \( f_{c1} \) and \( f_{c2} \) have been selected, the electrical length \( \theta_c \), in degrees, can be found from

\[
\theta_c = \frac{180}{\frac{f_{c2}}{f_{c1}} + 1}
\]  

(6.1)

At this point, from Table 4.1 the element values can be chosen. It has to be highlighted that the simulated behaviour of the filter differs from the theoretical specifications. That is, there are some effects that formulae do not take into account. Therefore, although the theoretical cutoff frequencies are \( f_{c1} = 3.1 \) GHz and \( f_{c2} = 10.6 \) GHz, respectively, the actual cutoff frequencies that result in the optimal bandpass filter are \( f_{c1} = 5 \) GHz and \( f_{c2} = 12.5 \) GHz. This specifications lead to an electrical length \( \theta_c = 51.43^\circ \). One can realize that there is no information about the element values for this value of \( \theta_c \). However, we can find them by interpolation from the other values presented in the table. As an illustration, for \( n = 6 \) and \( \theta_c = 51.43^\circ \), the element value \( y_1 \) is calculated by computing a straight line from the data in Table 4.1.

\[
m_1 = \frac{0.48096 - 0.35346}{35 - 30}
\]

(6.2)

\[
b_1 = 0.48096 - 35m_1
\]

(6.3)

\[
y_1(\theta_c) = m_1\theta_c + b_1
\]

(6.4)

\[
z_1 = \frac{50}{y_1}
\]

(6.5)
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

Where \( m_1 \) is the slope and \( b_1 \) the y-intercept value. In a similar way, the rest of elements can be found. With the values of all impedances and the electrical lengths of all connecting lines and stubs, it is straightforward to compute the line dimensions of the microstrip filter. The filter is realized in microstrip on a substrate with a relative dielectric constant \( \epsilon_r = 2.2 \) whose thickness is \( t = 1.57 \) mm. The filter is terminated by \( Z_0 = 50 \) ohms both ports. Using equations in section (4.2) the widths and physical lengths can be obtained.

The filter has been tested and optimized by full-wave EM simulation. Then, the filter has also been constructed for verifications, the dimensions of the bandpass filter are \( 5.7 \text{ cm} \times 2.6 \text{ cm} \). In Figure 6.6 both -the simulated and the real- response are plotted. Good agreement between measured and simulated results is observed between 2 GHz to 8 GHz. However, measured cutoff frequency \( f_c \) is lower than the simulated. Later on, the effects of this will be analyzed. The constructed filter prototype can be seen in Figure 6.7.

The final filter design is shown in Figure 6.8 and it is obtained by concatenating the realized bandpass filter and the shaping section previously described. The dimensions of this final structure are \( 9.4 \text{ cm} \times 3.5 \text{ cm} \). The calculated transmitted spectrum \( S(f) \) if this final design is used as \( H(f) \) in the block diagram 2.5 is shown in Figure 6.8.

The FCC mask is plotted in red, the emitted signal \( S(f) \) with the proposed filter in the system, in blue. And the output \( S(f) \) of the system without the filter, that corresponds to \( S(f) \) in the situation drawn in Figure 2.1, in green. For this simulation, the smoothed version of the Vivaldi antenna transfer function and the smoothed version of the pulse have been used.

From the simulations results, it can be evinced that making use of the proposed shaping filter, an improvement in the band occupancy is obtained at the same time that the mask is met all along the frequency range.

To verify the proposed solution, the prototype has been constructed and measured. The constructed prototype is shown in Figure 6.9. Also in Figure 6.9, the transmitted
signal $S(f)$, using the real measurements of the shaping filter, is plotted. These results have been obtained by inserting the measured data into equation (6.6). That is, the smoothed version of the FFT of the measured pulse created by the pulse generator, the measured filter response of the proposed microstrip structure and the measured antenna transfer function of the Vivaldi antenna.

$$G(f) \cdot H(f) \cdot A(f) \cdot 2\pi f = S(f) \quad (6.6)$$

As in the simulated case, in green we have the output without any filter, the transmitted spectrum is graphed in blue and the FCC mask in red.

**Analysis of the results in the frequency domain**

It is clear by observing the Figure 6.9 that the filter shapes the emitted spectrum $S(f)$ in such a way that makes it meet the FCC mask. Moreover, the spectrum occupies the
available spectrum more efficiently with the shaping filter than without it. However, the effect of the lower $f_{c2}$ of the bandpass filter in Figure 6.6 is now noticeable. The spectrum $S(f)$ is attenuated more than expected in the EM simulations. Consequently, a loss of efficiency occurs due to this fact. It will be partially fixed by replacing the bandpass filter by the stopband filter, as described in the next sections.

In order to compare the behaviour of the system with and without shaping filter, it is necessary to define a method to quantify the improvement. The way adopted in this work to measure the band occupancy is to compute the ratio between the power sent by the transmitting antenna and the total amount of power available inside the FCC mask. That is, observing Figure 6.9, the area below the blue graph from 3.1 GHz to 10.6 GHz divided by the area below the red line within the same frequency range. This ratio has to be computed in linear scale, since the power magnitude in logarithmic scale comprises from $-\infty$ to zero -for normalized magnitudes- and the result of the quotient would depend on the lower bound chosen to compute it. On the other hand, as the power is always positive in linear scale, its range is from zero to one if the power has been normalized.

The band occupancy achieved with the shaping filter amounts to 52.46%. For the sake of fairness, a comparison has to be made if both signals meet the FCC mask. It would make no sense to compare the band occupancy of the blue and green curves in Figure 6.9, since the blue one meets the FCC mask and the green does not. Therefore, two comparisons can be done. First, comparing the output $S(f)$, when the filter is in the system, to the output without the filter but with a power level as low as to respect the regulations. This situation is plotted in Figure 6.10. And second, comparing the output with the shaping filter and the output without a shaping filter but with a bandpass filter to make the emitted spectrum meet the mask. The band occupancy without any filter is 12.19% and with a bandpass filter is 38.57%. Now it can be reliably said that an improvement in the band occupancy exists if the shaping filter proposed in this section is used.
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

![Graph of power emission level vs. frequency]

**Bild 6.10:** Comparison between two outputs $S(f)$ meeting the FCC mask.

**Bild 6.11:** Phase response of the constructed filter.

**Bild 6.12:** Group delay of the constructed filter.

The phase response $\phi(f)$ of the constructed filter is plotted in Figure 6.11. It can be seen that it decreases linearly with the frequency. The filter group delay has also been analysed and it is plot in Figure 6.12. We can see that it is not constant since has a variation of more than 1 ns.

**Analysis of the results in the time domain**

Now, the time domain analysis presented in section 5.2 will be applied. The complex analytical response $h^+$ in the time domain [18] is computed by

$$h^+(k\Delta t) = \frac{1}{N\Delta t} \sum_{n=0}^{N-1} H^+(n\Delta f)e^{j(2\pi/N)kn}$$

(6.7)

where the measured data has been zero-padded from 0 GHz to 2 GHz. The result of (6.7) is a complex discrete-time function whose time resolution is $\Delta t = 1/f_{max}$, being
$f_{\text{max}} = 18 \text{ GHz}$. Then, the resulting time resolution is $\Delta t = 55.56$ picoseconds. If a finer interpolation is desired, the measured data can be complemented by zero padding from 18 GHz to 1018 GHz which leads to a time accuracy of $\Delta t = 0.98$ picoseconds. The magnitude $|h^+|$ is the impulse response envelope and the real part $\Re\{h^+\}$ is the time-discrete impulse response.

In this analysis, three elements will be compared. First, the filter itself. In previous section the filter transfer function has been shown and discussed. Now, the dual expression in the time domain, the impulse response, will be shown. Second, the transmitted time-domain waveform considering that the shaping filter is not present in the system. And third, the transmitted time-domain waveform taking into consideration the designed filter.

To obtain the envelope of the impulse response $|h^+|$ the inverse discrete Fourier transform (IDFT) in (6.7) is applied over the measured transfer function of the filter. The result is shown in Figure 6.13. To obtain the time-domain waveform of the transmitted signal, the antenna transfer function together with the $j2\pi f$ factor that models the time derivative, is converted using (6.7) to a time-domain function. Afterwards, and because the time domain measurements of the generated pulse is available, the convolution between the result of the IDFT and the pulse waveform is computed. The emitted waveforms, with and without filter are shown in Figures 6.15 and 6.14, respectively. All waveforms have been normalized to their maximum values. The difference between the time-domain waveform in Figure 6.14 and Figure 6.15 is the pulse width. The incorporation of the filter (Figure 6.13) into the system, spreads the transmitted signal proportionally to the impulse response width. All parameters to quantify the effect of the filter introduced in section 5.2 are summarized in Table 6.1.

**6.1.3 Shaping filter with bandstop filter**

The second method to meet the mask is to replace the bandpass filter with a bandstop filter. As Figure 6.5 reveals, the emitted spectrum at $f_{c2} = 10.6 \text{ GHz}$ is not meeting
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

The use of a bandpass filter has been addressed in last section and it has been shown that the real performance of the whole structure is degraded because of a too early cutoff rate.

To avoid this problem, a bandstop filter blocking signal from 2 to 3.1 GHz will be designed, simulated and tested. The filter used will be the bandstop structure discussed in section 4.3.2. The stubs consist of two sections with different impedance to avoid the periodicity of the regular bandstop stubs. The width of the narrow lines is designed to have an impedance of 140 ohms and the width of the wide sections to have an impedance of 35 ohms. With these impedance values, \( K = \frac{Z_2}{Z_1} = 0.25 \) is obtained, which is needed to compute the electrical lengths of the different sections. The lengths of the two stepped-impedance sections are obtained to extend the spurious stopbands beyond 10.6 GHz while attenuating the band comprised from 2 GHz to 3.1 GHz.

Its stopband response is shown in Figure 6.16.

In Figure 6.17 we can see the layout of the shaping filter designed in last section connected with the bandstop filter. It is also shown a simulation of the emitted spectrum \( S(f) \) when the designed filter is utilized in the system. This simulated emitted spectrum \( S(f) \) is obtained using the linear approximation of the antenna transfer function, the CST simulation of the filter response and the smoothed version of the generated pulse. At first glance, it can be noted that the output \( S(f) \) is satisfactorily meeting the FCC mask while providing good band occupancy. An advantage with respect the structure composed of the shaping and bandpass filter is clear. With the bandpass filter, via hole groundings and, hence, soldering is avoided which leads to a simpler fabrication process.
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

Bild 6.16: Bandstop filter response.

Bild 6.17: CST layout of the proposed filter (left) and the simulated $S(f)$ spectrum (right) when the filter is used.

It is also worth mentioning that the dimensions of the structure with the bandstop filter are slightly smaller than with the bandpass filter, that is $7.3 \text{ cm} \times 3.5 \text{ cm}$.

The structure in Figure 6.17 is constructed in order to test and verify it. The prototype is then measured to check the behaviour of the system. The plot in Figure 6.18 is obtained by inserting all measured data into equation (6.6). The green curve represents the output $S(f)$ without the shaping filter in the system, the blue curve is the transmitted spectrum with the designed structure. The FCC mask is plotted in red. $S(f)$ meets the FCC mask in all the frequency range but in $f = 2.26 \text{ GHz}$. Apart from that, it is clear that the efficiency increases when the structure is present in the system.

**Analysis of the results in the frequency domain**

The area below the FCC mask within the band of interest, from $3.1 \text{ GHz}$ to $10.6 \text{ GHz}$, occupied by the transmitted spectrum when the designed structure is within the system
6.1. FILTER FOR THE FCC MASK AND THE VIVALDI ANTENNA

Bild 6.18: Photograph of the constructed prototype (left) and the transmitted \( S(f) \) spectrum (right) when the filter is used.

is 52.40% of the total available spectrum. As in the last case, this area has been calculated in linear scale. This efficiency is 0.06% lower than with the bandpass. Regarding the spectrum occupancy we can now say that the increase of efficiency is around 14% with respect to the use of a bandpass and skipping the shaping filter.

Bild 6.19: Phase response of the constructed filter.

Bild 6.20: Group delay of the constructed filter.

In Figures 6.19 and 6.20, the phase of and the group delay of the filter from Figure 6.18 are plotted. Similarly as the previous filter, it exhibits non-constant group delay.

**Analysis of the results in the time domain**

Using the equations provided in section 5.2 the impulse response of the filter is obtained (see Figure 6.21). Likewise, the transmitted waveform \( s(t) \) with the pre-distortion filter \( h(t) \) is shown in Figure 6.22. The transmitted waveform without the filter has already
been shown in Figure 6.14. The time domain parameters are summarized in Table 6.2. The filter affects mostly in the ringing of the output. Comparing Table 6.1 and Table 6.2 we see that the filter using a stopband introduces less broadening than the filter using short-circuited stubs.

<table>
<thead>
<tr>
<th></th>
<th>FWHM</th>
<th>Ringing $\tau_{r,\alpha}=0.1$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter</td>
<td>0.1395 ns</td>
<td>0.6503 ns</td>
</tr>
<tr>
<td>$s(t)$ without pre-distortion filter</td>
<td>0.1591 ns</td>
<td>0.1768 ns</td>
</tr>
<tr>
<td>$s(t)$ with pre-distortion filter</td>
<td>0.1925 ns</td>
<td>0.7927 ns</td>
</tr>
</tbody>
</table>

Tabelle 6.2: Time domain parameters for the Vivaldi antenna and the FCC mask.

At this point, two filters to pre-distort the signal and compensate the effect of the Vivaldi antenna have been presented. It has been demonstrated that these shaping filters make the transmitted spectrum meet the FCC mask while achieving an improved adjustment to spectral regulations. The emission limits constraint considered up to now was the mask set by the Federal Communication Commission.

**6.2 Filter for the EC mask and the Vivaldi antenna**

**6.2.1 Shaping filter**

The goal will be to shape the spectrum $S(f)$ so that it is adjusted as efficiently possible to the mask established by the European Commission [16]. The shaping filter should pre-distort the generated pulse in such a way that the transmitting antenna radiates constant power in the 6 GHz to 8.5 GHz frequency range. For this purpose, the filter that has to be designed should approximate the shape in Figure 2.15. To obtain this shape, the linear approximation of the Vivaldi transfer function has been utilized as well as the smoothed version of the pulse. As it can be seen in Figure 6.23, the restrictions for the
EC mask are more severe than those for the FCC mask. The EC mask allows emissions in a narrower bandwidth and the required attenuation outside the band is higher than 25 dB. However, for the FCC mask, the needed attenuation is 10 dB. Consequently, a higher selective bandpass filter will be needed.

Bild 6.23: Emission limits for UWB indoor applications in Europe.

As the frequency band under study is considerably narrower than the FCC case, the use of a deep ripple is now feasible to implement. If we consider a lowpass filter with Chebyshev response with a certain cutoff frequency $f_c$, it will exhibit $n$ ripples within the band, since the Chebyshev response is equi-ripple in the passband. If the cutoff frequency $f_c$ and the degree of the filter $n$ are properly chosen, the lowpass filter might successfully approximate the target shape. This idea is illustrated in the example of Figure 6.24, taking $n = 3$, $f_c = 8.5$ GHz and a ripple of $\epsilon = 7.5$ dB.

Bild 6.24: Method to approximate the $S_{21}$ parameter of the pre-distortion filter.

The stepped-impedance filter discussed in section 4.4.1 has been selected for its simplicity. The first step is to choose the values for the low and high impedance transmission
6.2. FILTER FOR THE EC MASK AND THE VIVALDI ANTENNA

lines. Afterwards, the required ripple and the degree of the filter has to be chosen to compute the lowpass prototype element values. With all these and the formulae provided in section 4.4.1, the electrical lengths of each transmission line section are obtained. The last step is to convert electrical lengths and high and low impedance values to physical dimension of the microstrip lines using equations in section 4.2.

The filter response in Figure 2.15 suggests a ripple of around 10 dB which represents a problem. The software simulations of the structure resulting from all design equations for this filter are not satisfactory. This unusual filter specifications lead to unexpected filter responses. With all, the dimensions of the stepped-impedance structure may be changed in order to attenuate this drawback. By means of software optimization, a good shape approximation is obtained with but the deepest ripple level does not get even close to the required 10 dB. To solve this problem, the structure is duplicated and connected in series with a transmission line with characteristic impedance $Z_0 = 50$ ohms. After optimizing this new structure to correct concatenation mismatches, the layout of Figure 6.25 is obtained. The layout shows how the three-section stepped impedance filter structure is repeated twice to achieve a deeper ripple. The filter consists of an initial $Z_0=50$ ohms transmission line, followed by successive low, high and low impedance sections. Then, another 50-ohm transmission line is used to connect the other identical structure. As it can bee seen, the wide lines are the low impedance sections, while the narrow lines are the high impedance sections. The simulated response of the structure in Figure 6.25 is the red curve plotted in Figure 6.26. In blue, we have the ideal shaping filter. As it has been introduced, we can see that the red curve has a lowpass-like response with cutoff frequency $f_c = 8.5$ GHz and a ripple of around 10 dB which resembles the blue curve. The spectrum $S(f)$ plotted in Figure 6.27 has been obtained using the stepped-impedance circuit as $H(f)$ in the system (see Figure 2.5). Both, the red and green curve, have its maximum values within the band of interest set to 0 dB in order to compare the response in this band. The red curve corresponds to the output $S(f)$ without the

\[\text{Bild 6.25: CST layout of the proposed structure.}\]
6.2. FILTER FOR THE EC MASK AND THE VIVALDI ANTENNA

Bild 6.26: Comparison between ideal and simulated $S_{21}$ parameter.

Bild 6.27: Simulation of the emitted spectrum $S(f)$.

designed filter and the blue curve corresponds to $S(f)$ with the filter. The green line is indeed flatter from 6 GHz to 8.5 GHz than the red one although it clearly violates the mask outside this range.

The solution that is adopted is to use the bandpass filter presented in section 4.2.1. Several are the reasons to choose an open-circuited stub bandpass filter instead of using a bandpass filter made of short-circuited stubs. First, short-circuited stubs require soldering in order to connect the line with the ground plane. This can lead to degradation of the substrate, modifying the dielectric constant $\epsilon_r$ and hence, the guided wavelength $\lambda_g$. Soldering can also affect on the length of the stubs, changing their impedances. Now it is pertinent to point out the change of microstrip substrate with respect the case of the FCC mask. The new substrate with dielectric constant $\epsilon_r = 6.15$ with thickness $h = 0.635$ mm leads to the correct filter response. If this filter is implemented with the previous substrate, that is $\epsilon_r = 2.2$ with thickness $h = 1.57$ mm, the stubs widths become larger than their lengths. Consequently, the length of the bandpass filter increases to 20 cm and its response is completely unacceptable with the $S_{11}$ parameter above $-5$ dB in all the frequency band.

For these reasons, an open-circuited stubs bandpass filter with the new substrate is chosen. By observing Figure 6.27 it can be noted that considerable attenuation is needed at the edges of the band of interest. The required attenuation at $f_{c1} = 6$ GHz is approximately 30 dB, and about $-25$ dB at the upper cutoff frequency $f_{c2} = 8.5$ GHz. This specifications can be fulfilled with a bandpass filter of degree $n = 7$, that is, 7 open-circuited stubs which are $\lambda_g/2$ long with connecting lines that are $\lambda_g/4$ long, with a passband ripple $L_{Ar}=0.7$ dB and $h = 0.5$. The cutoff frequencies that result in the desired filter response are $f_{c1} = 6$ GHz and $f_{c1} = 9.2$. The dimension of all stubs and connecting lines are computed using equations provided in section 4.2.1 with the lowpass prototype elements values in section 3.2. The selected response is, as commented in section 3.2, Chebyshev. In order to justify the impossibility of implementing a Butterworth filter, it is convenient to obtain the stub widths of both the Chebyshev and Butterworth
filter. Using equations in sections 4.2.1 and 3.2, the line widths for the Butterworth response are of the order of few micrometers, while for the same filter specifications, the Chebyshev filter is composed of lines of the order of one millimeter. The final filter layout and its response are shown in Figure 6.28.

The filter response shows high selectivity and low insertion loss in the passband. However, as it has been pointed out in section 4.2.1, it has additional passbands at $f = 0$ and $f = 2f_0$ where $f_0$ is the midband frequency. It is certainly a problem since in those regions there is signal that has to be filtered out (Figure 6.27). To remove the spurious passband located in $f = 2f_0$ a lowpass filter will be used. The chosen technique will be the stepped-impedance structure discussed in section 4.4.1. Sufficient attenuation and cutoff rate are achieved with a filter of order $n = 7$. The layout of the lowpass filter and its $S_{21}$ parameter are found in Figure 6.29.

Next, the additional passband at $f_0=0$ has to be removed. As the antenna works from
2 GHz approximately and it does not radiate DC, only the frequency band from 2 GHz to 3.1 GHz has to be eliminated. Therefore, the appropriate filter should have a stopband response. Conventional stopband filters have spurious stopbands at odd multiples of the midband frequencies, so the first spurious stopband is at $3f_0$ where $f_0$ is the midband frequency of the band that has to be stopped. Unfortunately, this falls in the band that has to be optimized. This fact is illustrated with the response of a stopband filter designed to eliminate the passband at $f_0$. The layout of the conventional stopband filter and its response are provided in Figure 6.30.

The filter response is adequate at $f_0$ and could successfully attenuate the unwanted band located at around 3 GHz. Nevertheless, due to its periodical behaviour, the band comprised between 6 GHz and 8.5 GHz would be completely removed. Due to this fact it is convenient to use the stopband with extended passband presented in section 4.3.2.

Bild 6.31: CST layout of the proposed bandstop filter (left) and its simulated $S_{21}$ parameter (right).
In Figure 6.31 the layout of a extended passband bandstop filter is shown. Note that the stubs in Figure 6.30 have been replaced by stepped-impedance sections to break periodicity and avoid additional resonances. The stepped-impedance sections are separated $\lambda_g/4$, where $\lambda_g$ is the wavelength at the midband frequency $f_0$.

The bandstop filter is designed to attenuate the midband frequency $f_0 = 2.5$ GHz with four stepped-impedance sections as well as $f_0 = 3.5$ GHz with a single section. That is the reason why the last bandstop element is closer to the rest (Figure 6.32), because they are separated $\lambda_g/4$ where lambda depends on the midband frequency $f_0$. Furthermore, the first element has been placed up-side down to avoid coupling with the last stub from the bandpass filter. The high and low impedance sections have been designed to extend the first spurious stopband frequency five times the center frequency, that is $f_{s1}/f_0 = 5$, where $f_{s1}$ is the first spurious frequency.

### 6.2.2 Shaping filter with bandpass filter

Both the shaping filter and the microstrip structure that filters the unwanted signals out, to fulfill the EU mask, are designed. The following step is to connect them together and simulate the whole system with the layout, represented in Figure 6.32, added between the pulse generator and the transmitting Vivaldi antenna. The simulation of emitted spectrum $S(f)$ is plotted in Figure 6.32. The blue curve represents the emitted spectrum $S(f)$ when the pre-distortion filter is present in the system, and the green curve is the emitted spectrum is no filter were placed in the system. The simulation indicates that the concatenation of the bandstop, bandpass and lowpass filters works correctly and the resulting spectrum does not violate the EU mask. Moreover, the current emitted spectrum (in blue) is much flatter than before (in green). To perform these simulations, the smoothed version of the generated pulse and the smoothed version of the antenna have been used.
After having checked the good system performance with these simulations, the filter prototype has been constructed and tested. The photograph of the fabricated prototype is given in Figure 6.33. The total structure consists of the bandpass filter made of 7 open-circuited half-wavelength stubs, followed by the bandstop filter and finally, the shaping filter of the previous section. For this current situation, where the Vivaldi antenna is considered, the lowpass filter is not needed since the antenna attenuates sufficiently the high frequencies. Similarly, a bandstop of higher order \( n = 5 \) is required to provide enough attenuation at low frequencies. The dimensions of the structure are 14.4 cm × 2.8 cm.

After measuring the structure, the transmitted spectrum \( S(f) \) is plotted in Figure 6.33.

Analysis of the results in the frequency domain

Results reveal good agreement between simulations and experimental data. Similar as in the cases for the FCC mask, comparisons have to be made in order to quantify the improvement that the filter introduces to the system performance. The first comparison is between the area occupied by the emitted power \( S(f) \) within 6 GHz and 8.5 GHz when the filter is taken into consideration and the occupancy of the spectrum \( S(f) \) when the filter is not considered. In this last case, the emitted power should be such that the mask is fulfilled. An example of this situation is shown in Figure 6.34, where a very low power pulse is transmitted not to overpass the regulations. The band occupancy for this low power emission is 0.17%. If the designed shaping filter is considered, the band occupancy increases to 54.86%. The other comparison that has been done is if only a bandpass filter is used to select the 6-8.5 GHz frequency range without pre-distorting the transmitted pulse. An efficiency of 53.05% is obtained in this case. These results are poorer than the FCC case results due to the fact that the EC mask is much narrower than the FCC and the distortion that the antenna introduces in this band is indeed
less accentuated, which, minimizes the difference between the pre-distorted spectrum and the not pre-distorted spectrum. Moreover, as the transition bands are not ideal and the mask require high filter selectivity, the passband becomes narrower leading to a loss of efficiency. Despite these inconveniences, the best performance is achieved using the designed prototype and the efficiency might become better by increasing the bandwidth of the passband filter, since results in Figure 6.34 show that the passband is narrower than it was in simulations. To conclude the frequency domain analysis, the phase response and the group delay of the shaping filter is shown in Figures 6.35 and 6.36.

Bild 6.34: Comparison between two outputs $S(f)$ meeting the FCC mask.

Bild 6.35: Phase response of the constructed filter.

Bild 6.36: Group delay of the constructed filter.
Analysis of the results in the time domain

As described in 5.2, the frequency domain data is transformed to the time domain. Figure 6.37 shows the impulse response of the filter, Figure 6.38 shows the transmitted waveform $s(t)$ in the time domain without the effect of the shaping filter and, in Figure 6.39 the transmitted signal modified by the shaping filter is illustrated. These figures clearly reveal the influence of the filter over the transmitted signal $s(t)$. The duration of the ringing is increased in around 2.5 nanoseconds with respect the case without filter. Furthermore, the width of the envelope of the filter impulse response $|h^+|$ is four times wider than the impulse response envelope of the filters designed for the FCC mask. The same occurs for the duration of the ringing since the filter for the FCC case is wide in the frequency domain, which results in a very narrow impulse response in the time domain. On the contrary, the EC mask frequency range is much narrower which leads to a wider time domain response. All time domain parameters are summarized in Table 6.3.

<table>
<thead>
<tr>
<th></th>
<th>FWHM</th>
<th>Ringing $\tau_{r,\alpha=0.1}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter</td>
<td>0.7161 ns</td>
<td>2.6503 ns</td>
</tr>
<tr>
<td>$s(t)$ without pre-distortion filter</td>
<td>0.1591 ns</td>
<td>0.1768 ns</td>
</tr>
<tr>
<td>$s(t)$ with pre-distortion filter</td>
<td>0.6896 ns</td>
<td>2.6306 ns</td>
</tr>
</tbody>
</table>

Tabelle 6.3: Time domain parameters for the Vivaldi antenna and the EC mask.
6.2. FILTER FOR THE EC MASK AND THE VIVALDI ANTENNA

Bild 6.38: $s(t)$ without pre-distortion filter

Bild 6.39: $s(t)$ with pre-distortion filter
7 Filter design for the Bow-tie antenna

The current scenario at this moment is that depicted in Figure 2.5, where $A(f)$ is the idealized transfer function for the Bow-tie antenna. The transfer function of the Bow-tie antenna can be approximated with a constant line, as seen in Figures 7.1 and 7.2. This smoothed response is used in order to compute the shape of the required pre-distortion filter without taking into consideration the erratic and unpredictable behaviour of the antenna transfer function.

Bild 7.1: Transfer function in the main beam direction, co-polarized component for the FCC frequency range.

Bild 7.2: Transfer function in the main beam direction, co-polarized component for the EC frequency range.

7.1 Filter for the FCC mask and the Bow-tie antenna

7.1.1 Shaping filter

In this section, a pre-distortion microstrip filter will be designed to approximate the computed ideal filter illustrated in Figure 7.3. The considered emission mask is the FCC mask. Consequently, the transmitted spectrum should be as flat as possible within the frequency range comprised between 3.1 GHz and 10.6 GHz.

By comparing Figures 2.14 and 2.13 we can note that both curves are similar in shape but different in attenuation levels. In the case this section is dealing with, the minimum value for the $S_{21}$ parameter is about $-15$ dB which is about half the attenuation that the previous filter required. Nevertheless, the design procedure is analogous to the previous case. A modified parallel-coupled line bandpass filter is used again, for the reasons exposed in section 6.1. No conventional filter provides such a smooth transition band.
over such a wide bandwidth and no passband ripple can get that deep either. However, the current specifications suggest that the new degree of the filter should be lower. Indeed, making use of only one section of coupled lines and optimizing the spacing between lines and their length and impedance, the simulated response is acceptable. The final filter layout and its response are shown in Figure 7.4.

If the whole system (Figure 2.5) is simulated using the designed structure as \( H(f) \), the on-the-air spectrum \( S(f) \) results to be that plotted in Figure 7.5. To perform the simulation, the smoothed versions of the antenna transfer function and the pulse are used. The simulation illustrates that the emitted spectrum occupies a greater area than the emitted spectrum without the filter. However, it is not confined to the FCC mask so a bandpass filter is also required in this case. It should be underlined that, now, a bandstop filter cannot be used since attenuation below 3.1 GHz and beyond 10.6 is needed. Because of this, the same bandpass filter used in section 6.1.2 will be used here.
7.1. FILTER FOR THE FCC MASK AND THE BOW-TIE ANTENNA

Bild 7.5: Simulation of the emitted spectrum $S(f)$ using the plot of the Figure 7.4.

Bild 7.6: CST layout of the proposed filter (left) and the simulated $S(f)$ spectrum (right) when the filter is used.

7.1.2 Shaping filter with bandpass filter

The final layout, that is, the shaping filter together with the bandpass filter, is shown in Figure 7.6. Proceeding as in the last section, the structure is first simulated before constructing. This simulation is plotted in Figure 7.6 and it reveals good band occupancy while lying below the regulation limits in all frequencies. Therefore, design is now ready to be built. The photograph of the constructed prototype and the simulation of the system with the real measurements of the filter are provided in Figure 7.7. The dimensions of the structure are $7.7 \text{ cm} \times 3.5 \text{ cm}$.

Analysis of the results in the frequency domain

A comparison between the performance of the system with and without pre-distortion filter is given in Figure 7.7. The emitted spectrum $S(f)$, in blue, is above the green line
(S(f) without filter) in most frequency range and fully respects the emission mask. The band occupancy with the proposed filter is 47.59%. If the filter is not used, the output would violate the mask (see Figure 7.7 green curve). Therefore, the emission power should be adjusted in order to make the emitted spectrum fulfill the regulations. If this is done, the band occupancy from 3.1 GHz to 10.6 GHz is 12.84%. An intermediate solution can also be adopted. If only a bandpass filter (without shaping filter) is used to filter out the spectrum outside the band of interest without compensating in any way the effect of the transmitting antenna an efficiency of 40.65% is obtained. Thus, the best results are achieved with the incorporation of the shaping filter in the system. 

As demonstrated in Figures 7.8 and 7.9, the phase response is not linear and the group delay is not flat. The impact of the variation of the group delay can be seen in the time domain waveforms plotted in Figures 7.11 and 7.12, where the level of the ringing and
7.2 FILTER FOR THE EC MASK AND THE BOW-TIE ANTENNA

FWHM is increased when the pre-distortion filter is used with respect to the case where only the pulse generator and the transmitting antenna are in the system.

![Impulse response and envelope](image)

Bild 7.10: Filter impulse response.

**Analysis of the results in the time domain**

Converting the frequency domain measurements to time domain functions with equation (6.7), the filter impulse response and the transmitted wave $s(t)$ are found. In Figure 7.10 the impulse response of the filter and its envelope are shown. The effect of adding a filter in the Ultra-Wideband transmission end can be noticed by comparing Figure 7.11 and Figure 7.12. As it was expected and similarly to the previous cases, the filter broadens the width of the transmitted pulse as well as the duration of the ringing.

7.2 Filter for the EC mask and the Bow-tie antenna

**7.2.1 Shaping filter**

The last filter design is addressed in this section. The forth case consists of considering the Bow-tie antenna as the transmitting antenna $A(f)$ in the block diagram of Figure 2.5. The power emission limitation is again the mask set by the European Commission. Therefore, the goal is to transmit a flat spectrum over the band starting at 6 GHz and finishing at 8.5 GHz, while respecting the power regulations in all the frequency range.

First, the design of the shaping filter to make $S(f)$ as flat as possible within the band of interest will be considered, regardless of the behaviour of the system outside this band. The target is to find a microstrip structure whose $S_{21}$ parameter resembles the $S_{21}$ parameter plotted in Figure 2.16. This case is very similar to the case of the Vivaldi antenna with the only difference that now less attenuation is required. The starting attenuation value at $f = 6$ GHz is around 5 dB when it was 10 dB for the Vivaldi antenna.
Taking advantage of these similarities the same kind of structure is adopted. However, the dimensions of the stepped-impedance sections as well as the connecting lines should be modified to have the required attenuation at the lower edge. Unlike the previous case and owing to the fact that the difference between attenuation levels at both edges is low (around 5 dB), the stepped-impedance structure does not shape the passband of the bandpass filter sufficiently to have a good approximation of the $S_{21}$ parameter in Figure 2.16. In other words, if a shaping filter with the ideal response within the band of interest is obtained and connected to a bandpass filter, the resulting structure will not have this perfect response anymore. The solution is to design a shaping filter with a greater difference between the highest and lowest attenuation levels than the computed ideal response says. This is illustrated in Figure 7.13.
7.2.2 Shaping filter with bandpass filter

The bandpass used for this case differs from the one used in the previous section. Unlike the Vivaldi antenna, the Bow-tie antenna does not attenuate high frequencies since its transfer function can be approximated with a constant line. Consequently, the lowpass filter is needed in this case. At the same time, the required attenuation levels in the low frequencies are less demanding than before, so a bandstop filter of lower order is used now. The design procedure is exactly the same as in section 6.2 with the difference that the last element of the bandstop filter is not necessary. So, in this case, the bandstop filter is composed of only four stepped-impedance stubs. The layout of the resulting filter is given in Figure 7.14 along with the simulation of the transmitted spectrum $S(f)$. Now, it is demonstrated that although the $S_{21}$ parameter of the shaping filter is very different from what it should be (see Figure 7.13), the transmitted spectrum $S(f)$ is quite flat in the band from 6 to 8.5 GHz. Also, good compliance with the FCC mask is obtained.

![Image of CST layout and simulated spectrum](image)

Bild 7.14: CST layout of the proposed filter (left) and the simulated $S(f)$ spectrum (right) when the filter is used.

The prototype is then constructed and measured. The dimensions of the final microstrip structure are 14.5 cm × 2.8 cm. The use of the measured data to compute the expression in (6.6) leads to the graph in Figure 7.15, where the blue curve is the radiated spectrum $S(f)$ when the designed circuit is in the system, the green curve is the output $S(f)$ when this circuit is not used and, in red, the EC mask is provided.

**Analysis of the results in the frequency domain**

The EC mask is respected in all frequency range by the radiated spectrum $S(f)$. The comparison methods are the same that have been carried out along all this work. The band efficiency when a pulse without any kind of pre-distortion is used is 0.21%. This low efficiency is due to the fact that very low power has to be sent not to violate the EC mask. This situation has been represented in Figure 6.34. The occupancy achieved with
the pre-distortion filter is 55.09%. If the designed shaping filter is replaced by a bandpass filter that performs no pre-distortion over the generated pulse, the emitted spectrum $S(f)$ meets the mask but not as effectively as with the designed pre-distortion filter. The band occupancy in this situation is 53.92%. Similarly to the last case, the improvement is moderate and the reasons are also alike. The narrowness of the band together with the smooth shape that has to be made, makes it difficult to correct with accuracy. That is, the power difference, without pre-distortion, between the two frequency edges is of the order of the fluctuations of the antenna response which are difficult to compensate. However, observing Figure 7.15 some ideas to improve the results show up. In the frequency range comprised from 2 GHz to 6.5 GHz, the difference between the mask and the radiated power $S(f)$ is around 15 dB which suggest that the order of the stopband filter might be decreased or even eliminated from the structure which could lead to reducing the dimensions of the pre-distortion filter as well as a possible passband enhancement. The same idea could apply for the lowpass filter.

In Figures 7.16 and 7.17, the phase response and the group delay are plotted. The filter does not exhibit flat group delay and a peak is found in 6.2 GHz due to an abrupt fall of the phase.

**Analysis of the results in the time domain**

In Figure 7.18 the filter impulse response is plotted in the time domain. The pulse width can be explained for its relatively narrow frequency response which entails relatively large pulse width. This width causes that the effect of the filter over the emitted waveform $s(t)$ is more noticeable, as we can note by comparing Figures 7.19 and 7.20. All time domain parameters are provided in table 7.2.
7.2. FILTER FOR THE EC MASK AND THE BOW-TIE ANTENNA

Bild 7.16: Phase response of the constructed filter.

Bild 7.17: Group delay of the constructed filter.

Bild 7.18: Filter impulse response.

<table>
<thead>
<tr>
<th></th>
<th>FWHM</th>
<th>Ringing $\tau_{r,\alpha=0.1}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter</td>
<td>0.1778 ns</td>
<td>0.6385 ns</td>
</tr>
<tr>
<td>$s(t)$ without pre-distortion filter</td>
<td>0.1837 ns</td>
<td>0.1719 ns</td>
</tr>
<tr>
<td>$s(t)$ with pre-distortion filter</td>
<td>0.2348 ns</td>
<td>0.4764 ns</td>
</tr>
</tbody>
</table>

Tabelle 7.1: Time domain parameters for the Bow-tie antenna and the FCC mask.

<table>
<thead>
<tr>
<th></th>
<th>FWHM</th>
<th>Ringing $\tau_{r,\alpha=0.1}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Filter</td>
<td>0.6955 ns</td>
<td>1.056 ns</td>
</tr>
<tr>
<td>$s(t)$ without pre-distortion filter</td>
<td>0.1837 ns</td>
<td>0.1719 ns</td>
</tr>
<tr>
<td>$s(t)$ with pre-distortion filter</td>
<td>0.7230 ns</td>
<td>1.3507 ns</td>
</tr>
</tbody>
</table>

Tabelle 7.2: Time domain parameters for the Bow-tie antenna and the EC mask.
Bild 7.19: $s(t)$ without pre-distortion filter.

Bild 7.20: $s(t)$ with pre-distortion filter.
8 Conclusions

The goal of this thesis was to find a microwave passive circuit to pre-distort the signal in such a way that the radiated spectrum adapted to the given spectral mask as much as possible. That is, maximum power emission was wanted while respecting the emission constraints.

Four different microstrip structures have been designed to improve the spectral usage for four different scenarios. First, the Vivaldi antenna has been considered as transmitting element and the FCC regulations as the spectral mask. Second, the Bow-tie antenna has been the antenna under study and the FCC mask the limitations to fulfill. Third and fourth case have dealt with the same pair of antennas but taking into consideration the European spectral mask instead of the American mask.

The use of traditional bandpass filters has been considered as a possible solution for the radiated spectrum to meet the given spectral mask. But, as bandpass filters only select a band of interest without modifying in any way the shape of the input signal, the emitted spectrum is severely affected by the antenna transfer function and by the generated pulse spectrum itself. Therefore, the mask is not properly fulfilled since maximum power within the mask is not emitted.

To solve this inefficiency, four shaping filters that pre-distort the signal have been designed, constructed and measured. It has been proved that a better mask fulfillment is obtained using these microstrip structures. The pre-distortion is carried out with filters to compensate the effect of the non-ideal antennas and to amplify the spectrum in frequency range where the power of the pulse is low. With all constructed structures, both masks have been fully respected and the spectral usage has been improved in all cases.

To perform this work, a theoretical analysis of the transmitting end of the UWB system has been performed. MATLAB® has been utilized to compute all needed expressions and to obtain the required pre-distortion filters. The design has been carried out using passive microstrip filter such as lowpass filters, bandpass filters and stopband filters. Further research and optimization methods have been necessary to accomplish the goal. All microstrip structures presented in this text have been designed, simulated and optimized using CST MICROWAVE STUDIO® 2008.

It is noteworthy that best improvements have been obtained for the case of the FCC mask, because of its huge bandwidth (the non-ideal antenna behavior is more noticeable and there is more room for improvement) and for its low demanding specifications compared to the European mask which calls for 30 and 25 dB attenuation at each edge. On the other hand, the FCC mask only asks for 10 dB attenuation with respect to its maximum emission level.

As well as the frequency domain analysis, the expressions of the pre-distortion filter impulse responses, in time domain, have been provided and analyzed. The impact of
adding a shaping filter to the system has been a broadening of the transmitted pulse. This broadening has been more significant in the cases where the European mask was considered. The reason behind this fact is because in these cases, the transmitted spectrum was confined in a narrower bandwidth than the case of the FCC mask. And, the narrower the signal bandwidth is, the wider the impulse response becomes.

Finally, measurements have revealed that certain circuit elements (that were essential according to the simulations) were actually redundant or not completely necessary. The removal of those elements could lead to a less degraded passband response and to a reduction in the dimensions of the microstrip structure. In order to achieve a higher reduction in the filter dimensions, a different technique may be used. The low temperature co-fired ceramic (LTCC) technology can design circuits on several layers [34], [35]. With this method, resistors, capacitors and inductors can be printed on different layers which leads to a shrinkage of the filter.
Literaturverzeichnis


[17] *Picosecond Pulse Labs PSPL 3600 User Manual*


