ESCUELA TÉCNICA SUPERIOR DE INGENIERÍA DE TELECOMUNICACIÓN
UNIVERSIDAD POLITÉCNICA DE CARTAGENA

Proyecto Fin de Carrera

Rediseño y optimización de un receptor de alta frecuencia para sondeo de canales MIMO

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Resumen

Se denomina MIMO, *Multiple Input Multiple Output*, a aquellos sistemas de múltiples antenas que permiten en determinadas circunstancias de propagación mejorar la capacidad del canal radio mediante el aprovechamiento de la estructura de multicamino inherente al canal de radio móbil con respecto al caso SISO, *Simple Input Simple Output*. Dada la complejidad y el relativo poco tiempo desde que surgieron este tipo de sistemas, el canal MIMO es relativamente desconocido y muchas de las ventajas que pueden obtenerse de este tipo de sistemas están todavía por descubrir.

Por esta razón la Universidad de Hannover investiga para obtener un alto conocimiento de los sistemas MIMO. Anteriormente a este proyecto ya se dio el primer paso verificando por medio de simulaciones el potencial que ofrece este tipo de estructura de antenas.

Este proyecto comienza el segundo paso en la consecución de dicho objetivo. Este paso consiste en el diseño de un receptor suficientemente robusto que elimina las desventajas de diferentes prototipos diseñados en esta universidad. Este diseño tiene en cuenta todos los problemas que pueden surgir en este tipo de canales y trata de proporcionar una señal a la salida de alta calidad.

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REDISEÑO Y OPTIMIZACIÓN DE UN RECEPTOR DE ALTA FRECUENCIA PARA SONDEO DE CANALES MIMO

1. Introducción

El siguiente documento resume el proyecto fin de carrera realizado por Alfonso Rosagro Escámez, alumno de la Universidad Politécnica de Cartagena, durante su periodo en Hanóver (Alemania) como estudiante Erasmus. El proyecto original está redactado en inglés y se hará referencia a él en determinadas ocasiones.

El instituto de tecnología de alta frecuencia y sistemas de radio está trabajando para satisfacer la creciente demanda de anchos de banda de transmisión y velocidad de datos a través de circuitos transmisores y receptores. Este trabajo se centra en explotar la estructura de multi-camino que ofrece el canal de comunicación inalámbrico para incrementar la velocidad de datos mediante sistemas MIMO y la calidad de la señal recibida mediante diversidad. Se denomina MIMO, Multiple Input Multiple Output, a aquellos sistemas de múltiples antenas que permiten en determinadas circunstancias de propagación mejorar la capacidad del canal radio con respecto al caso de transmisiones con una sola antena en transmisión y recepción. Concretamente dicho departamento ha desarrollado un conjunto de antenas planares de banda ancha capaz de adaptar estos esquemas de transmisión ya que son capaces de proporcionar caminos independientes de transmisión. El potencial que ofrecen estas estructuras de antenas con respecto a la comunicación MIMO ha sido verificado por medio de simulaciones basadas en análisis de prestaciones.

Lo que se pretende en este momento es la verificación de estas simulaciones por medio de medidas reales en un canal inalámbrico en el interior de edificios. Para ello, se ha desarrollado un concepto básico de un sistema MIMO y se ha comenzado con la implementación y realización de este sistema.

Concretamente, el proyecto que pretende ser explicado en este documento consiste en llevar a cabo el sondeo de un canal MIMO en el interior de un edificio. Para ello, la principal tarea consiste en el diseño y optimización de un receptor especial de alta frequencia, así como su fabricación e integración con los equipos del laboratorio de dicho departamento con el fin de llevar a cabo dicho sondeo del canal. El receptor diseñado durante este proyecto se muestra en la siguiente imagen:

Figure 1.1: Receptor para aplicaciones MIMO

El concepto desarrollado contiene también una gran cantidad de procesado de señal. Por ello, el proyecto ha sido realizado en paralelo con otro proyecto que se encarga de la sincronización de códigos, recuperación de la portadora transmitida así como estimación de fase y frecuencia y ciertos parámetros del canal MIMO.

2. Canal de comunicaciones inalámbrico

Para llevar a cabo el diseño de un receptor de alta frecuencia en canales inalámbricos es necesario el previo conocimiento de los efectos que el canal inalámbrico produce en una comunicación ya que éste establece varios límites tanto a la señal transmitida como a los requerimientos del receptor. El canal inalámbrico ofrece también posibilidades a la transmisión que están todavía por explotar. La utilización de sistemas MIMO trata de beneficiarse de estas propiedades. Una visión general sobre el canal inalámbrico y los sistemas MIMO es dada a continuación.

Los efectos del canal inalámbrico se pueden distinguir claramente en la señal recibida. Un ejemplo de una señal recibida puede se muestra en las siguientes figuras:

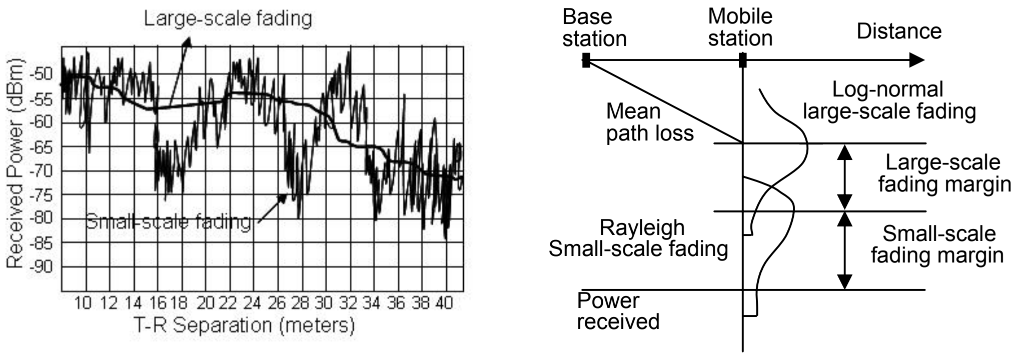


Figura 2.1: Efectos del canal inalámbrico en la señal recibida

Dos contribuciones a la señal recibida pueden ser observadas en la figura. Una debida a los llamados efectos de gran escala (large scale fading) y otra debido a los efectos de pequeña escala (small scale fading).

Los efectos de gran escala representan la atenuación de la potencia media de la señal debido a grandes desplazamientos o grandes obstáculos en el medio de propagación. Estos efectos se pueden apreciar en la señal recibida como dos contribuciones, una atenuación media debido a la distancia y una variación sobre la media de carácter aleatorio en función del canal concreto medido. En la siguiente figura se pueden ver los elementos que producen estos efectos (izquierda) y el método que se utiliza para estimar estos efectos (derecha).

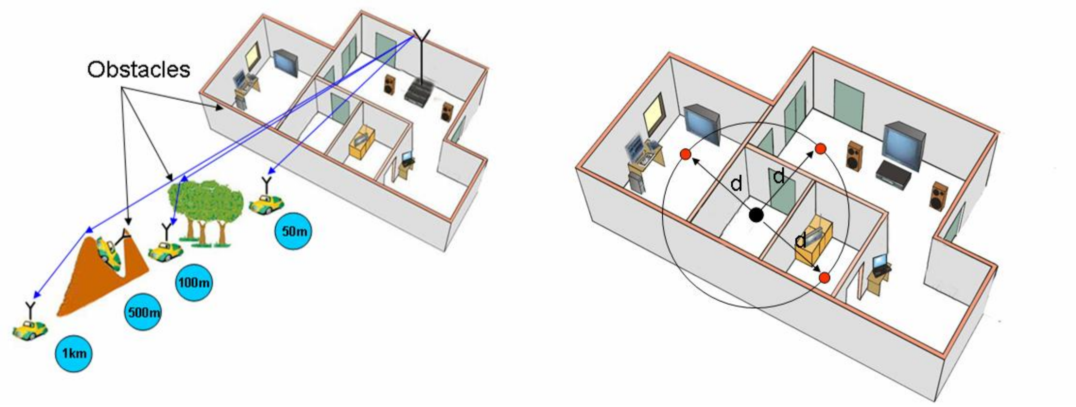


Figura 2.2: Efectos de gran escala

El método representado en la figura se conoce como el modelo de Okumura-Hata y determina la atenuación media que sufre la señal a una determinada distancia debido a los efectos de gran escala, mediante una media de las atenuaciones en diferentes puntos situados a la misma distancia. La expresión definida por el modelo de Okumura-Hata para calular la atenuación media en función de la distancia viene dado por la siguiente expresión:

$$\overline{L}_p = \frac{d}{d_0}^{n} \xrightarrow{DECIBELS} \overline{L}_p(dB) = L_p(d_0)(dB) + 10 \cdot n \cdot \log(d/d_0) \qquad (2.1)$$

$L_p(d_0)$ son las pérdidas a una distancia d_0. La distancia d_0 corresponde a un punto localizado en el campo lejano de la antena y frecuentemente se toma 1 metro para canales en el interior de edificios. La fórmula aplicada para calcular estas pérdidas es normalmente la expresión de las pérdidas en espacio libre:

$$L_s = \left(\frac{4 \cdot \pi \cdot d}{\lambda}\right)^2 \xrightarrow[DECIBELS]{d=1} L_s(dB) = 20 \cdot \log(4 \cdot \pi / 0.3) + 20 \cdot \log(f(GHz) \qquad (2.2)$$
$$= 32.44dB + 20 \cdot \log(f(GHz)$$

El valor del exponente n depende de la frecuencia, el tamaño de las antenas y el medio de propagación. En espacio libre, n es aproximadamente igual a 2, pero en presencia de obstáculos este valor es mayor.

De la fórmula dada por Okumura Hata se puede observar que al incrementar la separación entre las antenas, el incremento de las pérdidas es muy grande en los primeros metros pero conforme va aumentandose la distancia, las pérdidas van incrementando en menor cantidad.

El valor medio dado por la fórmula de Okumura Hata no es adecuado para caracterizar un canal concreto. En este caso se introduce en la expresión una variable aleatoria con una distribución log-normal resultando en la siguiente expresión:

$$\overline{L}_P = \frac{d}{d_0}^n \xrightarrow{DECIBELS} \overline{L}_P(dB) = L_s(d_0)(dB) + 10 \cdot n \cdot \log(d/d_0) + X_s \tag{2.3}$$

Valores de X_s entre 6 y 10 dB basados en diferentes medidas realizadas son muy utilizados en ingeniería.

Los efectos de pequeña escala representan cambios dramáticos en la amplitud y fase de la señal como resultado de pequeños obstáculos en el medio de propagación o pequeños desplazamientos que producen una variación en la separación entre receptor y transmisor.

Estos efectos son modelados habitualmente con una función de densidad de probabilidad de Rayleigh que describe la envolvente de la señal recibida en una transmisión multi-camino (un símbolo transmitido llega al receptor por diferentes caminos debido a reflexiones en el medio) en el peor caso, es decir, cuando la principal componente o el rayo directo no se recibe.

Los efectos de pequeña escala están caracterizados por dos mecanismos, la dispersión de la señal y la variación del canal con el tiempo.

La dispersion de la señal se debe a los diferentes caminos que sigue hasta llegar al receptor y que hacen que el canal se pueda ver como una serie de pulsos en el dominio del tiempo. La siguiente figura ilustra el efecto de la dispersión en el dominio del tiempo y de la frecuencia.

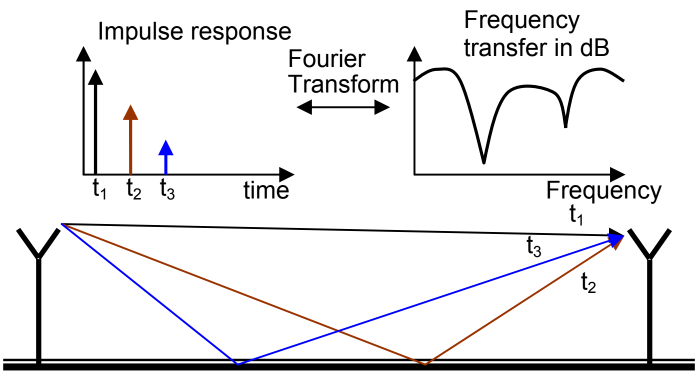


Figura 2.3: Caracterización de los efectos de pequeña escala: dispersión de la señal

La dispersión de la señal está caracterizada por el máximo tiempo de retardo de un símbolo transmitido. El máximo tiempo de retardo es el intervalo desde que se recibe la primera componente del símbolo transmitido hasta que se recibe la última componente con significante energía. Este tiempo impone un límite inferior al tiempo de símbolo transmitido, ya que, si el máximo tiempo de retardo es mayor que la duración del símbolo, los componentes de un símbolo llegarán cuando el siguiente símbolo es recibido, provocando pérdida de información. El efecto que se produce es pérdida de información de forma semejante a la que provocaría un filtro en una señal cuyo ancho de banda es mayor que la del filtro.

Por otro lado, la variación del canal está caracterizada por el tiempo de coherencia (Tc). El tiempo de coherencia se refiere a la propia naturaleza cambiante del medio debido a pequeños movimientos de objetos en el canal o de las antenas transmisoras o receptoras. Concretamente define el tiempo en el que el canal permanece aproximadamente constante o, en otras palabras, el tiempo a partir del cual el canal decorrela las señales que le llegan.

En la siguiente figura se muestra la función de autocorrelación AC(Δt) que define como la respuesta al impulso del canal decorrela las señales a lo largo del tiempo. Para valores de AC(Δt) mayores que cero, el canal se comporta de forma correlada, por tanto, Tc define el rango de valores en los que AC(Δt) no es cero.

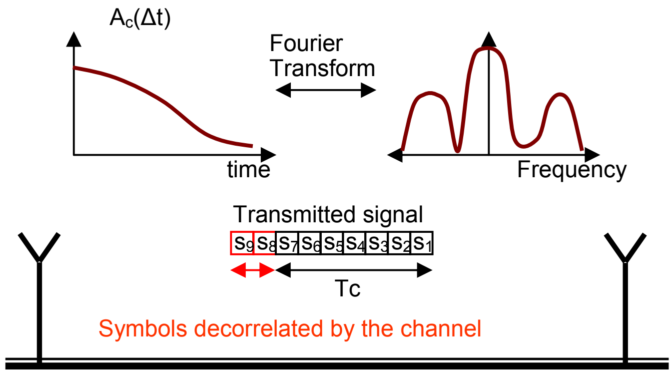


Figura 2.4: Caracterización de los efectos de pequeña escala: dispersión de la señal: variación del canal con el tiempo

Para valores de Δt mayores que Tc la función de autocorrelación es cero y por tanto las señales separadas en tiempo una distancia mayor del tiempo de coherencia estarán incorreladas tras pasar por el canal. Por tanto, el tiempo de coherencia impone un límite superior al tiempo de símbolo transmitido, ya que, si el tiempo de símbolo hace que la transmisión de una determinada secuencia de símbolos es mayor que el tiempo de coherencia, símbolos recibidos pertenecientes a una misma transmisión llegarán incorrelados.

Como se comentó en la introducción, una forma de explotar el canal inalámbrico es utilizar sistemas MIMO. Un sistema MIMO es un enlace entre multiples antenas transmisoras y múltiples antenas receptoras. Un sistema MIMO genérico puede verse en la siguiente figura:

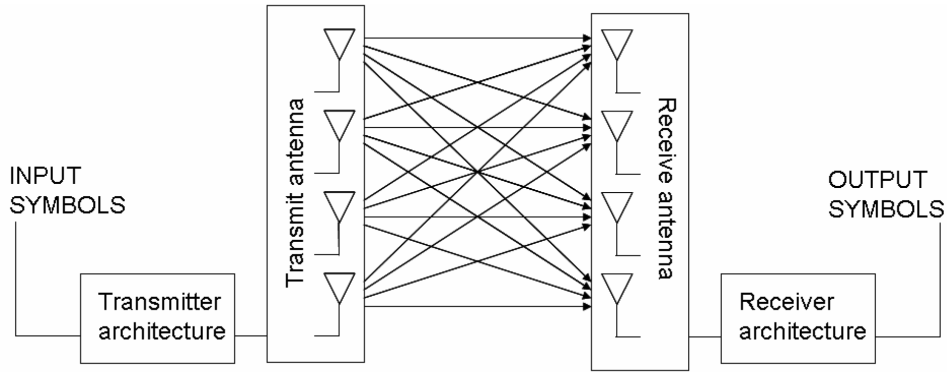


Figura 2.5: Sistema MIMO genérico

El principio de funcionamiento es la transmisión en paralelo de información a través de un array de antenas transmisoras y receptoras situadas a una distancia fija unas de otras. Con esta configuración del array de antenas se consigue una multiplexación en espacio. Esto le da al canal una gran complejidad, pero consigue obtener del canal una capacidad mucho mayor que la que se obtendría con un sistema SISO (Single input-Single output).

Mediante este esquema de transmisión cada antena receptora recibirá contribuciones de cada una de las antenas transmisoras. Con el fin de separar las diferentes señales recibidas, las señales transmitidas en cada antenas se correlan en transmisión y se decorrelan en recepción consiguiendo la separación de las señales recibidas. Para la correlación de las señales en transmisión se multiplica una portadora por una señal digital diferente de pseudoruido para cada una de las antenas. Previamente la señal de pseudo-ruido se modula utilizando el esquema de modulación deseado. Tras la multiplicación estas señales son transmitidas al canal y recibidas por cada una de las antenas receptoras. Cada una de las antenas receptoras conoce el código de cada una de las señales transmitidas y por tanto, mediante cierto procesado de la señal, es capaz de separarlas.

A la vez que se consigue mayor capacidad del canal mediante el sistema MIMO también es posible conseguir mayor calidad de señal aplicando técnicas de diversidad. Es decir, cada antena le aplica diferentes regímenes a las señales de llegada, lo que facilita el separarlas en recepción. Un ejemplo de diversidad es la diversidad espacial. Cada elemento del array de antenas receptoras se coloca en diferente posición de forma que recibe con una diferente fase a cada señal de llegada basándose en su ángulo de llegada.

3. Arquitecturas de receptores

Una vez conocidas los efectos del canal y posibles sistemas para aprovechar sus propiedades, el diseño propiamente dicho del receptor puede comenzarse.

La primera fase del diseño consiste en elegir su arquitectura. Existen muchos tipos de arquitecturas posibles y cada una tiene sus ventajas y desventajas. En la siguiente figura se pueden ver algunas de las posibles arquitecturas.

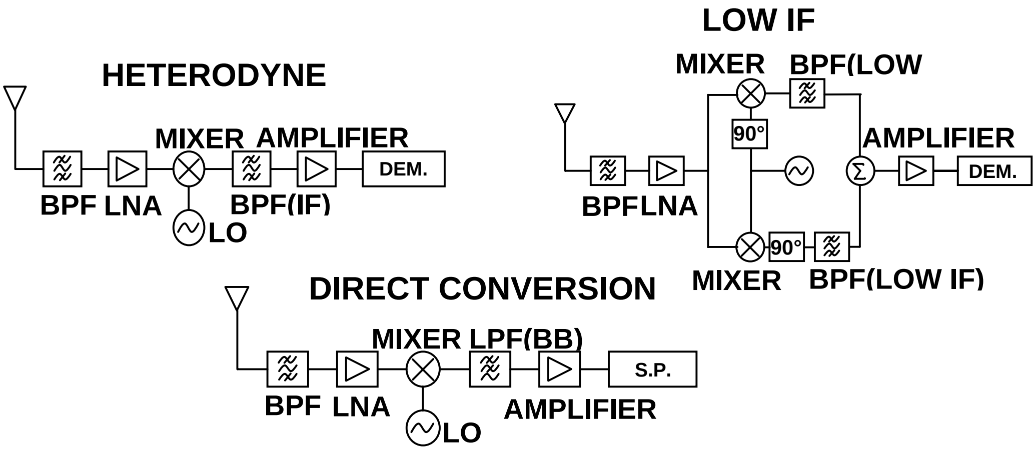


Figura 3.1: Arquitecturas de receptores

La arquitectura adoptada en este proyecto es la arquitectura de baja frecuencia intermedia ("low IF architecture"). El principio de funcionamiento de esta arquitectura consiste en convertir la señal de radio frecuencia recibida por la antena a una frecuencia intermedia muy pequeña. La razón de esta elección es que dicha arquitectura evita muchos de los problemas que tienen las demás. Por ejemplo evita el offset de continua (DC offset) que afecta a la arquitectura de conversión directa ("direct conversion").Este problema se evita ya que al trabajar con la señal a una frecuencia intermedia, ninguna información irá a la frecuencia cero y por tanto, esta frecuencia se puede filtrar sin pérdida de información. Otro problema que evita es la necesidad de filtros externos, ya que al trabajar con una frecuencia fija y muy pequeña, estos filtros se pueden diseñar integrados. Otros problemas que evita esta arquitectura pueden ser vistos en el documento redactado en inglés.

El principal inconveniente de la arquitectura de baja frecuencia intermedia es el problema del rechazo de la frecuencia imagen. Este problema se puede ver gráficamente en la siguiente gráfica:

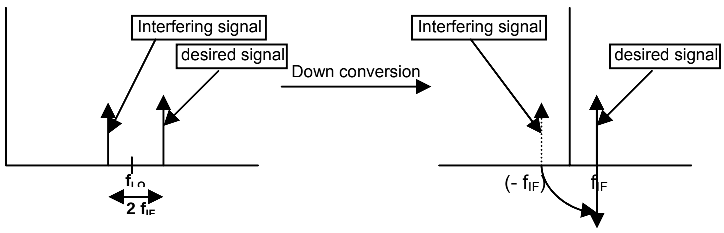


Figura 3.2: Problema de la frecuencia imagen (image frequency)

El problema del rechazo de la imagen consiste en lo siguiente. Cuando la señal de radio frecuencia es recibida por la antena, un filtro de preselección puede usarse para eliminar las componentes frecuenciales fuera de la banda de interés. Sin embargo, debido a que la frecuencia intermedia utilizada por esta arquitectura es muy pequeña, interferencias situadas por debajo de la frecuencia de la señal deseada a una distancia de dos veces la frecuencia intermedia (llamadas señales imagen) requerirían de filtros de frecuencia muy complicados para eliminarla. Por tanto un filtro de preselección no puede ser usado para eliminar esta frecuencia.

Por otro lado, cuando la señal de radiofrecuencia (señal deseada) es convertida a la frecuencia intermedia, la señal imagen es convertida a menos la frecuencia intermedia. Debido a las propiedades del espectro en frecuencia, esta señal imagen aparece a la misma frecuencia que la señal deseada como puede verse en la figura 3.2. Por tanto, un filtro de frecuencia tampoco puede ser usado para el rechazo de la señal imagen.

En este caso, lo que hace la arquitectura de baja frecuencia intermedia es eliminar la señal imagen jugando con la fase de la señal. Hay varias formas de realizar esto. Algunas de ellas son explicadas en el documento redactado en inglés. Para rechazar la imagen en este proyecto se utiliza un filtro de polifase de dos etapas. Este filtro es una de las partes fundamentales de este trabajo y su funcionamiento será explicado en un capítulo posterior.

4. Funcionamiento del receptor

Una breve visión general de cómo trabajan los distintos componentes de esta arquitectura es dada a continuación. La siguiente figura representa el circuito receptor diseñado.

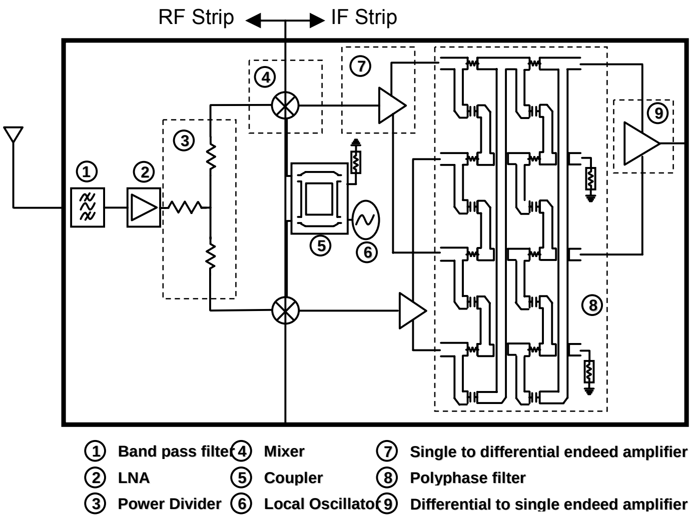


Figura 4.1: Circuito del receptor

En la primera etapa, un filtro paso banda (1) es colocado para eliminar las componentes frecuenciales fuera de la banda deseada. El segundo componente después del filtro paso banda es un amplificador de bajo ruido (2) que amplifica la señal recibida que ha sido debilitada durante la transmisión. El siguiente componente es un divisor de potencia (3) que divide la señal en dos caminos. Cada una de estas señales llega a un mezclador (4) que la convierte a una frecuencia más pequeña, llamada frecuencia intermedia. El valor de la frecuencia intermedia depende de la frecuencia del oscilador local (6). La señal generada en el oscilador local va a la entrada de un acoplador (7) que introduce una diferencia de fase entre sus dos salidas de 90º. Estos 90º son necesarios para obtener la componente en fase y cuadratura en los dos caminos respectivamente. El siguiente bloque al que le llega la señal es un amplificador diferencial (7) que convierte la señal simple que le llega en una señal diferencial. Esto se produce en ambos caminos de manera que el filtro de polifase (8) recibe a su entrada dos señales diferenciales. Éste trabaja con las señales como se explicará en un capítulo posterior y devuelve a la salida otras señales diferenciales en las que se ha eliminado la frecuencia imagen presente a su entrada. El último bloque (9) toma una de las salidas diferenciales del filtro polifase y las convierte otra vez a señal simple para que pueda ser tratada por los instrumentos que trabajan sólo con este tipo de señales.

Cuando la señal llega a la salida de este último bloque, ésta tiene que ser almacenada. Para ello se ha utilizado en este proyecto el analizador de señales vectoriales modelo 89600. El analizador vectorial de señales se muestra en la siguiente figura:

Figura 4.2: Analizador vectorial de señales

Del analizador vectorial de señales dependen las características necesarias en la señal recibida para que esta pueda ser reconocida y demodulada. La información de este instrumento puede ser encontrada en el documento redactado en inglés.

5. Parámetros del receptor

El parámetro más importante del receptor es su rango dinámico. Éste se puede definir como el rango entre la máxima y la mínima potencia de señal que puede ser tratada por el receptor sin significante ruido o significantes impedimentos. Por tanto el receptor se debe diseñar para que la señal deseada recibida esté dentro de ese rango. En la siguiente figura se puede ver el receptor diseñado y los factores de los que depende dicho rango:

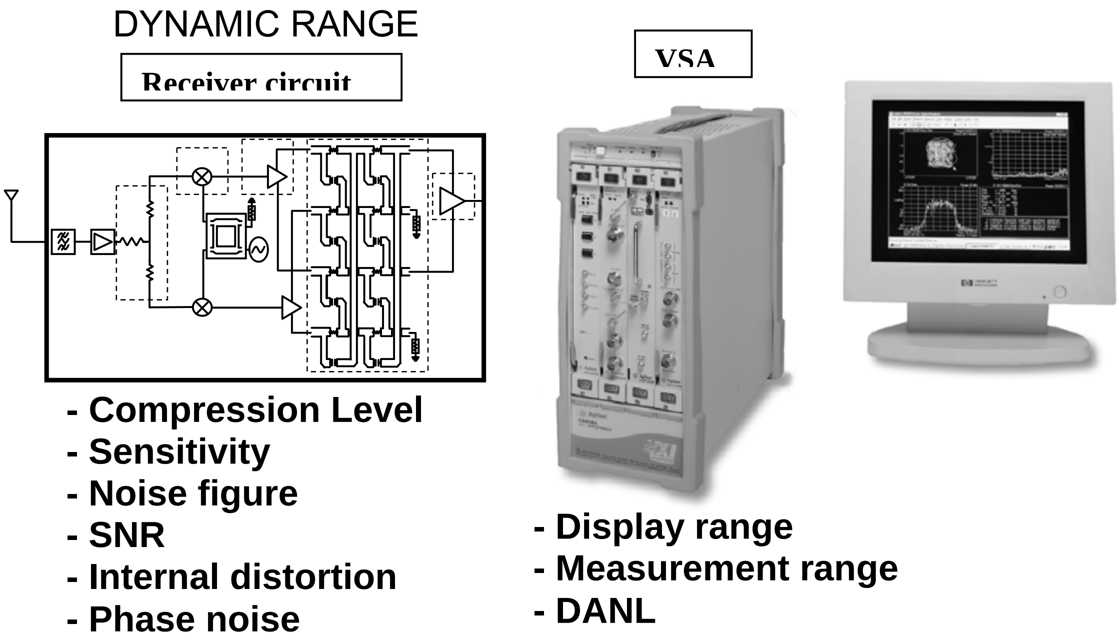


Figura 5.1: Parámetros que determinan el rango dinámico

Cada uno de estos factores puede determinar el rango dinámico del receptor dependiendo de las medidas que se estén realizando o de las condiciones en las que se realicen las medidas. Una detallada información sobre cada uno de estos parámetros puede verse en el documento redactado en inglés al que se refiere este resumen.

Los parámetros que mayor atención merecen a la hora de diseñar un receptor son la sensibilidad del instrumento y la relación señal a ruido necesaria para demodular la señal, ya que el receptor debe estar diseñado para proporcionar un nivel de potencia de la señal deseada por encima del valor de la sensitividad y manteniendo ese margen contra el ruido. También merece especial atención la figura de ruido que relaciona los parámetros anteriores con la potencia de la señal recibida que se desea remodular y su valor define las prestaciones del receptor.

Los demás parámetros definen los niveles de potencia con los que puede tratar el receptor y también deben ser conocidos puesto que pueden hacer explicar porqué ciertas señales no pueden ser recibidas.

Otros parámetros interesantes se refieren a la presencia de interferencias. Por ejemplo, para el caso del rechazo de la imagen explicado anterioremente, está el parámetro “rango de rechazo de la imagen” que se calcula a partir de los errores de amplitud y fase introducidos por el circuito. Y para el caso de la presencia de dos señales cercanas interferentes que debido a productos de intermodulación acaban en la misma frecuencia que la señal deseada, se usa el parámetro IP3. Una amplia explicación sobre este parámetro es dada en el documento redactado en inglés.

Una vez conocidos los parámetros que hay que tener en cuenta en el receptor así como los factores limitantes, se puede empezar a definir ciertos valores concretos para el receptor. Por ejemplo, conocidos los tiempos limitantes que imponen los factores de pequeña escala explicados anteriormente, pueden ser elegidos los parámetros del instrumento que almacena los datos de forma que se pueda transmitir una determinada señal para una determinada aplicación. En el caso de este proyecto, se definen los valores del analizador vectorial de señales para poder recibir una señal suficientemente larga (con muchos símbolos) de forma que facilite la labor al algoritmo de reconocimiento de señales utilizado en el proyecto realizado en paralelo con este.

Los efectos de gran escala sin embargo van a determinar la ganancia que necesitará el circuito receptor para que cuando la señal llegue al analizador vectorial de señales, tenga la suficiente potencia como para ser reconocida.

Estos valores así como los diferentes parámetros del receptor explicados anteriormente pueden ser consultados en el documento redactado en inglés.

Un esquema con los valores de los diferentes parámetros calculados y las relaciones entre ellos se muestra a continuación:

Time specifications

Max time record length 10 ms.

Signal transmitted:
- Symbol length = 1000 symbols
- BW = 300 kHz
- Signal time length = 6,66 msec
- Symbol time length = 6,66 µsec

VSA parameters:
- Span = 2 MHz → Sample rate = 5,12 MSa/sec
- RBW = 300 kHz

→ Samples per symbol = 34,09 samples
→ Time record length = 6,82 msec

Power specification

VSA parameters:
- Span = 2 MHz → Sample rate = 5,12 MSa/sec
- RBW = 300 kHz
- DANL = -151 dBm/Hz

RBW, DANL → DANL = -96,23 dBm

SNR_{min} = 20 dBm

$P_{min\ input}$ = -90 dBm

DANL, SNR_{min}, $P_{min\ input}$ → $Gain_{min}$ = 13,77

RF receiver NF = 10,77 dB

Wireless channel:
- Noisefloor = -174 dBm/Hz
- Channel losses:
  - d = 1m → $\bar{L}_p$(dB) = 40,223 ± X_s
  - d = 10m → $\bar{L}_p$(dB) = 70,223 ± X_s
  - d = 50m → $\bar{L}_p$(dB) = 91,192 ± X_s

$P_{transmitted}$ = 0 dBm

→ Pin = -50
→ Pin = -80
→ Pin = -101

Interference = -40 dBm → $IP3_{max}$ = 10 dBm

IRR_{min} = 25 dB

6. Diseño receptor y optimización del receptor

Hasta este punto se ha tratado de dar un resumen de los conceptos teóricos aplicados en el proyecto. A partir de ahora serán explicadas las características más importantes de los elementos escogidos para el receptor una vez escogidos teniendo en cuenta todos estos conceptos teóricos explicados anteriormente. En el documento en inglés también puede encontrarse información sobre el substrato escogido y cómo afecta este al diseño.

Antes de hablar de componentes concretos, todos ellos cumplen una característica, la simetricalidad. Todos los elementos han sido escogidos con la condición necesaria de que sean simétricos. Esto es debido a que el receptor trabaja con las señales eliminándolas o amplificándolas por medio de sumarlas y restarlas. Por lo tanto errores en la simetricalidad de los componentes producen una importante degradación de la señal recibida.

De los componentes elegidos, el divisor de potencia, el acoplador y sobre todo el filtro de polifase son los componentes del receptor no prefabricados, es decir, diseñados desde cero a base de líneas de transmisión y diferentes componentes pasivos. Por tanto estos han sido los componentes que han pasado por el proceso de simulación, fabricación y comprobación. Los demás componentes, es decir, amplificadores de bajo ruido, mezcladores y amplificadores diferenciales han sido encargados a diferentes fabricantes. Por tanto, con estos últimos componentes se parte de unas especificaciones dadas por el fabricante que han sido escogidas suficientemente altas para cubrir las especificaciones deseadas para el receptor. Por lo tanto se les ha añadido redes externas de elementos pasivos para proporcionarles las condiciones óptimas según el fabricante y han sido directamente testados para comprobar sus especificaciones. Las características de estos últimos componentes comparados con las características deseadas del receptor pueden ser vistas en el documento redactado en inglés.

Los primeros tres elementos nombrados son descritos a continuación brevemente.

Divisor de potencia. En este proyecto se ha optado por un divisor de potencia usando líneas microstrip y resistencias. A pesar de las pérdidas que implican las resistencias, éstas se han incluido para conseguir la adaptación de todos los puertos. Cuando se usa una resistencia en cada uno de los brazos del divisor de valor $Z_0/3$, se puede demostrar matemáticamente que todos los puertos están adaptados. Esta demostración matemática puede ser encontrada en el documento redactado en inglés al que se refiere este documento.

Por tanto la red diseñada es como se muestra en la siguiente figura:

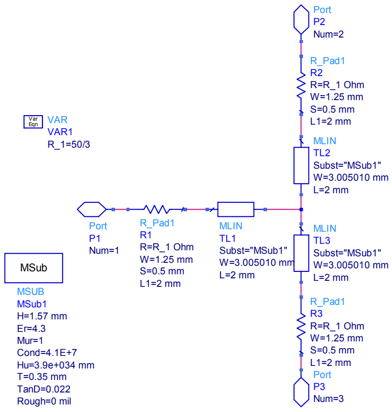


Figura 6.1 Divisor de potencia

Este esquemático ha sido creado en el programa ADS para su simulación y comprobación de sus propiedades.

A parte de las pérdidas introducidas y la adaptación de los puertos, otra propiedad a comprobar es su simetricalidad, o lo que es lo mismo, la reciprocidad de su matriz de parámetros S.

Para comprobar estas tres propiedades, este componente ha sido simulado con el programa ADS y los resultados obtenidos han sido los siguientes.

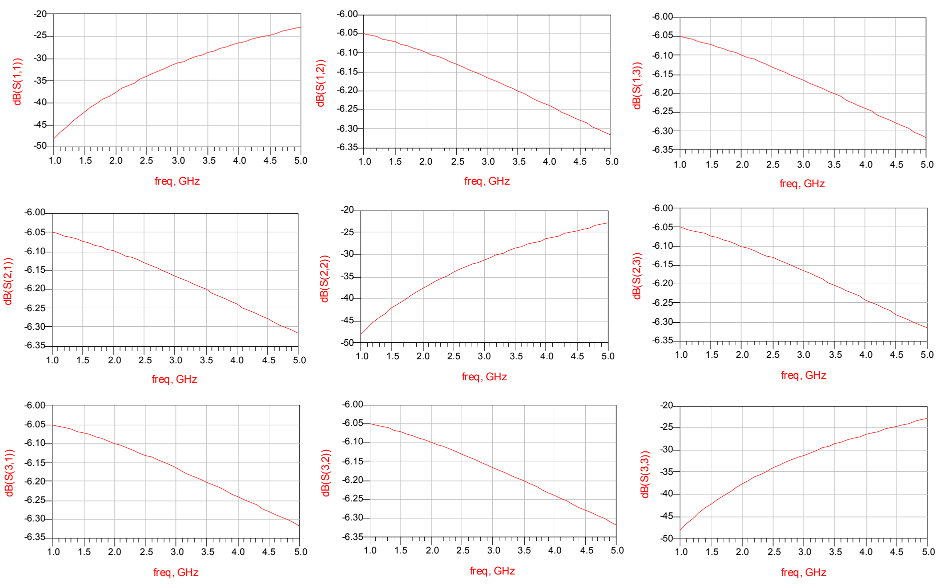


Figure 6.2: Parámetros S del divisor de potencia

Como se puede ver en la figura anterior S11, S22, y S33 presentan valores muy pequeños, por lo que se pueden considerar todos los puertos adaptados. De los demás parámetros S se deduce que el circuito introduce unas pérdidas de aproximadamente 6 dB (3 dB debido a las resistencias y 3 dB debido a la división de potencia en dos caminos) y que la red es perfectamente simétrica, ya que introduce los mismos efectos en todos los puertos.

El siguiente componente es el más importante del diseño y por tanto una explicación un poco más detallada que para el resto de componentes es dada a continuación. Se trata de un filtro de polifase con el que se pretende resolver el problema del rechazo de la imagen. Como se explicó anteriormente, al ser convertida la señal a frecuencia intermedia con el mezclador, una posible interferencia situada a dos veces la frecuencia intermedia será convertida a la misma frecuencia que la señal deseada por las propiedad de espejo de frecuencia que tiene el espectro de la señal. Sin embargo, la señal imagen de frecuencia tendrá una fase diferente que la señal deseada. Esto lo aprovecha el filtro de polifase para eliminarla.

El funcionamiento de este filtro es sencillo y puede ser explicado más fácilmente utilizando señales sinusoidales a su entrada y explicando que ocurre con ellas.

En las siguientes líneas se explica brevemente este funcionamiento utilizando señales sinusoidales.

Como explicado anteriormente, la señal deseada y la señal imagen estarán situadas antes del mezclador a $(\omega_{LO} + \omega_{IF})$ y $(\omega_{LO} - \omega_{IF})$ respectivamente. Por tanto se pueden representar estas señales de la siguiente forma:

Señal deseada → $$S(t) = As \cdot \sin\left[(\omega_{LO} + \omega_{IF})t\right] \quad (6.1)$$

Señal imagen → $$M(t) = Am \cdot \sin\left[(\omega_{LO} - \omega_{IF})t + \Delta\delta\right] \quad (6.2)$$

Por tanto estas señales irán por mezcladas en los dos caminos en los que se divide la señal tras el divisor de potencia.

Después de pasar por el mezclador, es decir, después de ser multiplicadas por una señal de la siguiente forma, $G \cdot \sin(\omega_{LO} \cdot t)$en un camino (camino I), y noventa grados adelantada $G \cdot \cos(\omega_{LO} \cdot t)$ en el otro camino (camino Q) y teniendo en cuenta las relaciones trigonométricas entre senos y cosenos que pueden ser consultadas en la bibliografía, la señales en ambos caminos quedan de la siguiente forma:

Path I

$$\frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta) \quad (6.3)$$

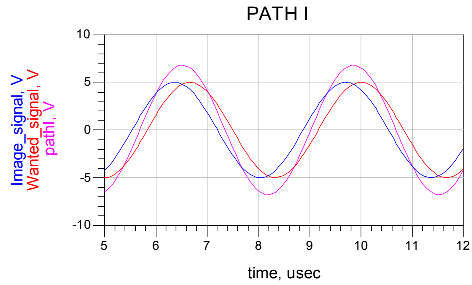


Figure 6.3: Señales en el camino I

Path Q:

$$\frac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)-\frac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta) \tag{6.4}$$

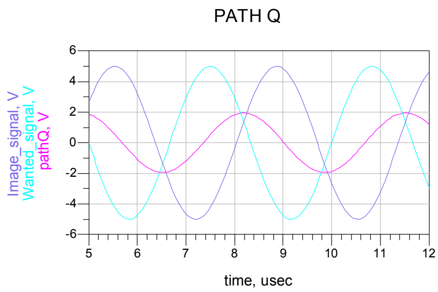


Figure 6.4: Señales en el camino Q

Estas señales son convertidas a señales diferenciales para poder ser tratadas por el filtro polifase. Para ello se utiliza un amplificador diferencial en cada camino, obteniendo las siguientes señales:

Differential signal (path I)

$$\begin{cases}\frac{G}{2}\cdot As\cdot\cos(\omega_{IF}\cdot t)+\frac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta)\\ -\frac{G}{2}\cdot As\cdot\cos(\omega_{IF}\cdot t)-\frac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta)\end{cases} \tag{6.5}$$

Differential signal (path Q)

$$\begin{cases}\frac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)-\frac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta)\\ -\frac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)+\frac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta)\end{cases} \tag{6.6}$$

El funcionamiento del filtro de polifase es de tal modo que las componentes de la señal a su entrada debidas a la señal interferente, se suman en contrafase y por tanto se eliminan y las componentes debidas a la señal deseada se suman en fase y por tanto se amplifican. Esto lo consigue el filtro de polifase de la siguiente forma explicada utilizando la siguiente representación del filtro de polifase:

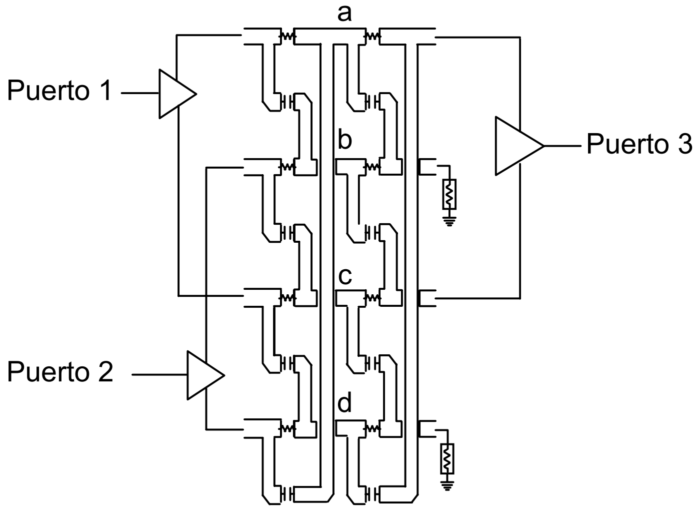


Figure 6.5: Filtro de polifase

La señal introducida en el puerto 1 llega al punto “a” del circuito con un desfase de -45º. Mientras que la señal introducida en el puerto 2 llega al puerto “a” con un desfase de -135º. En los puntos “b”, “c” y “d” los defases son (+45, -45), (-225, +45) y (-135, -225) respectivamente.

Debido a que el efecto es el mismo en todos los puntos nombrados anteriormente, sólo la señal en uno de esos puntos se muestra a continuación:

$$\text{POINT “a”} \begin{cases} \frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t - 135) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta - 135) + \\ + \frac{G}{2} \cdot As \cdot \sin(\omega_{IF} \cdot t - 45) - \frac{G}{2} \cdot Am \cdot \sin(\omega_{IF} \cdot t + \Delta\delta - 45) \\ = G \cdot As \cdot \cos(\omega_{IF} \cdot t - 135) \end{cases} \tag{6.7}$$

Por tanto, en estos puntos sólo se encontrará la señal querida.

Este funcionamiento del filtro polifase sólo se da a una determinada frecuencia. Esta frecuencia depende de la constante temporal que definen sus componentes (condensadores C y resistencias R). Concretamente, la frecuencia a la que se produce este funcionamiento es:

Diseñando el filtro polifase a esta frecuencia la función de transferencia que observaríamos viene dada en la siguiente figura:

$$f = \frac{1}{2 \cdot \pi \cdot R \cdot C}$$

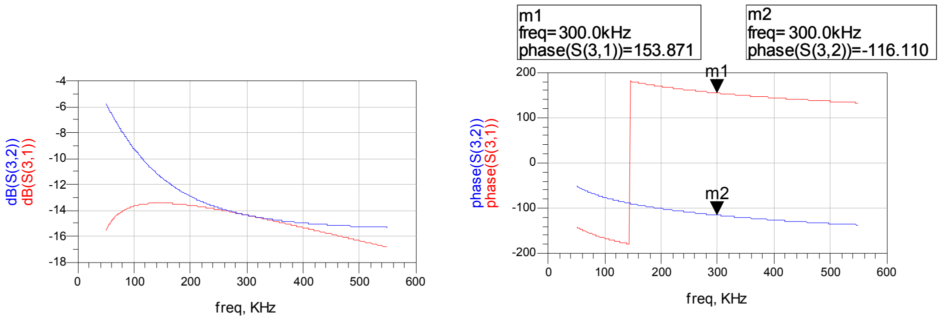


Figura 6.6:Parámetros ese del filtro de polifase diseñado para una frecuencia de operación

De la representación de la magnitud en dB de los parámetros S de transferencia se observa que a la frecuencia de diseño las dos salidas tendrán exactamente la misma amplitud y por lo tanto, debido al comportamiento del filtro polifase explicado anteriormente, el rechazo e la imagen será total.

Debido a que la señal que se transmite en este proyecto es una señal con un determinado ancho de banda, un filtro con una sola constante RC no es útil. En este caso se utiliza un filtro polifase de varias etapas, utilizando una constante de tiempo distinta en cada una de las etapas. El filtro polifase utilizado en este proyecto es el de la figura … Se aprecia que tiene dos etapas. Cada una de las etapas tiene unos valores de C diferentes cambiando la constante de tiempo de forma que se consiga un mayor ancho de banda de operación del polifase. Para los valores de C utilizados la función de transferencia obtenida es la siguiente:

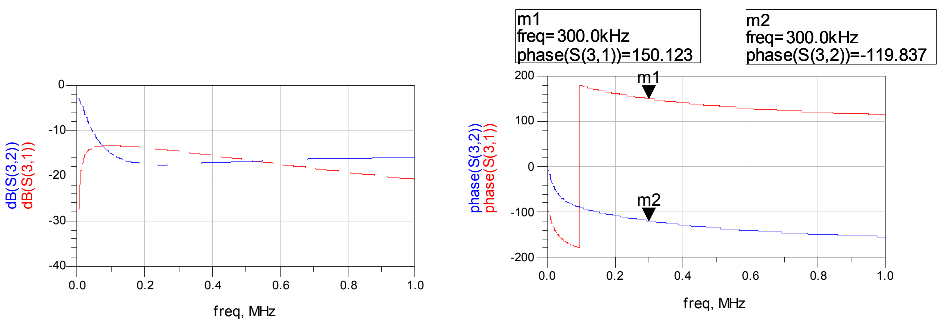


Figura 6.7:Parámetros ese del filtro de polifase diseñado para un rango de frecuencias

Se observa que se tendrá un pequeño error de amplitud a la frecuencia de diseño, sin embargo, para un determinado ancho de banda se mantiene un error de amplitud pequeño. Los valores exactos utilizados para la constante de tiempo y el porqué de su utilización, así como más detalles sobre el diseño del filtro de polifase pueden verse en la versión escrita en lengua inglesa al que se refiere este resumen.

El último elemento explicado es el acoplador con líneas de transmisión. Como se mencionó cuando se describió el funcionamiento del circuito, se necesitan dos señales de la misma amplitud y con un desfase de 90 º a la frecuencia del oscilador local para conseguir mediante los dos mezcladores las dos componentes de la señal recibida (componentes en fase y cuadratura). Debido a que la frecuencia del oscilador local es bastante alta, el desfase de 90ª se podría conseguir utilizando una única línea de transmisión. Con el fin de buscar simetricalidad para el receptor, se utiliza un componente con esta propiedad. Este componente es un acoplador de líneas de transmisión y puede verse en la siguiente figura:

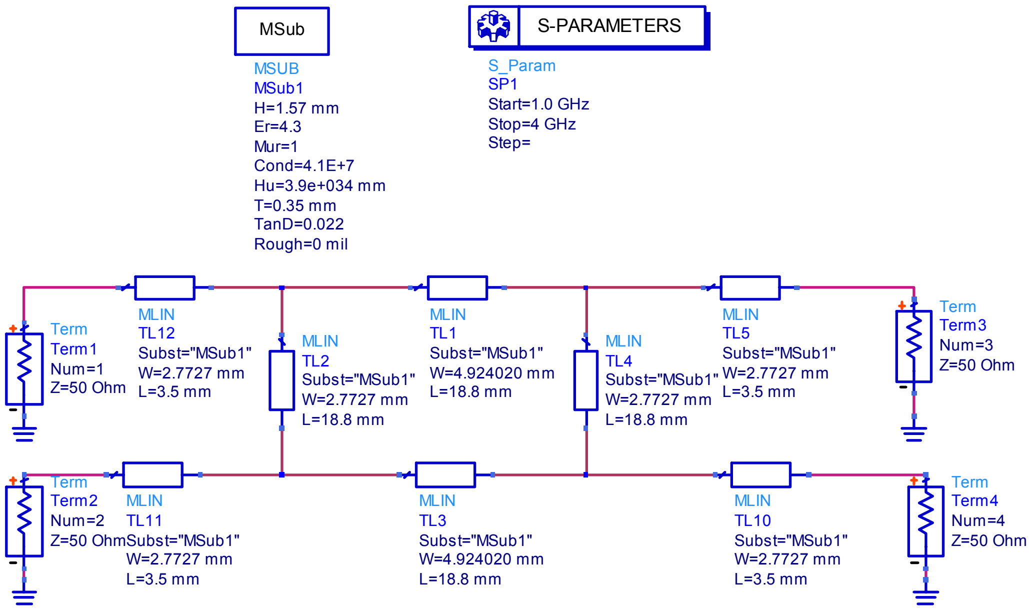


Figura 6.8: Esquemático de un acoplador de líneas de transmisión

Con el fin de conseguir las propiedades buscadas, cada uno de los brazos del acoplador tiene que tener una longitud de un cuarto de la longitud de onda de la señal y una impedancia igual a la impedancia a la que debe estar adaptado el circuito en los brazos laterales (Z0) y $Z_0/\sqrt{2}$ en los brazos superior e inferior.

Estos valores han sido calculados en el documento redactado en lengua inglesa al que se refiere este documento.

Para los valores calculados los resultados obtenidos mediante las simulaciones de estos parámetros son los siguientes:

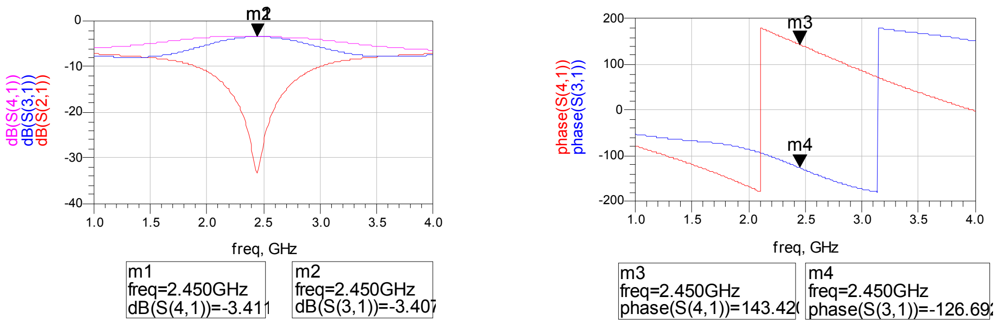


Figura 6.9: Parámetros S del acoplador

De la figura se observa que cuando se utiliza uno de los puertos como entrada, en dos de los puertos de salida se obtiene, a la frecuencia de trabajo, dicho desfase entre las señales y con un error de amplitud mínimo y el otro puerto está aislado.

Una vez se conocen todos los componentes utilizados en el diseño, el siguiente paso consiste en conocer los instrumentos las condiciones necesarias para la realización de loas medidas así como los montajes de los componentes y las medidas reales realizadas.

Los instrumentos que se han utilizado para las medidas de los componentes del receptor son los siguientes:

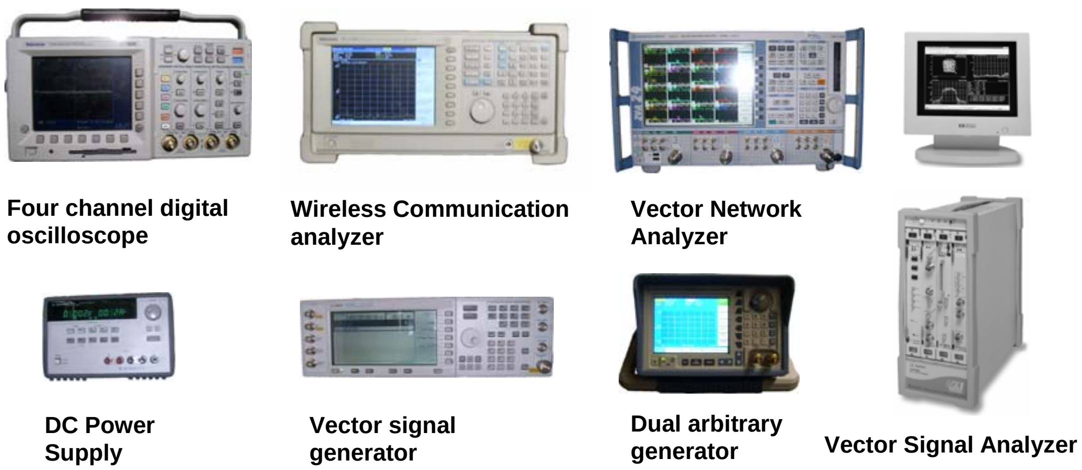


Figura 6.10: Instrumentación utilizada durante el proyecto

En cuanto a las condiciones en que se han realizado las medidas, lo que se ha buscado es el aislamiento de los componentes bajo prueba. Para ello se han utilizados unas cajas epeciales. Estas cajas y las dimensiones de cada una de ellas pueden verse en la siguiente figura:

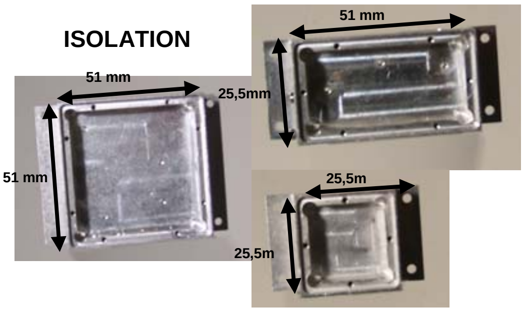


Figura 6.11: cajas para aislamiento de los circuitos

Sin embargo no todos los componentes han podido ser medidos aislados en estas cajas por el excesivo consumo de tiempo y gastos que ello supone. Para medir cada uno de los instrumentos de forma aislada se ha utilizado un banco de pruebas que permite la medida de los componentes de forma rápida aunque expuesto a degradación de la señal debido a la falta de aislamiento de los materiales. El banco de pruebas utilizado puede verse en la siguiente figura:

Figura 6.12: Bancos de prueba

Finalmente los componentes se han dividido en bloques y cada uno de estos bloques se ha introducido en una de estas cajas protectoras. Los cuatro componentes en que se han dividido pueden verse en la siguiente figura:

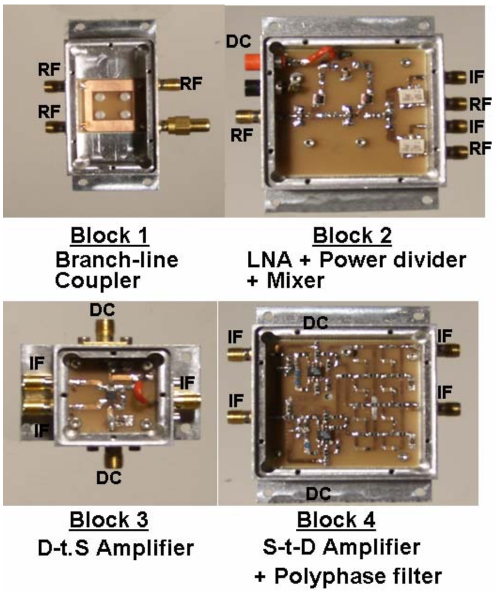


Figura 6.13: Bloques del receptor

Cada uno de estos bloques se ha medido por separado y después uniéndolos para comprobar el funcionamiento del receptor completo. Medidas mediante la transmisión de señales en diferentes puntos utilizando antenas bipolares para transmitir y recibir las señales han sido también realizadas comprobando el buen funcionamiento del receptor.

Los diferentes montajes realizados para la realización de las medidas de cada uno de los bloques pueden verse en el documento en inglés. Todas las medidas de estos componentes han dado los resultados esperados con muy pequeños errores que permiten la optimización del receptor.

Conclusiones y líneas futuras

En este trabajo se ha presentado el diseño desde el nivel más alto hasta el nivel más bajo de un receptor de radio frecuencia para un sistema MIMO. Primero se ha estudiado el medio en el que se transmiten las señales y cómo este medio afecta al diseño del receptor. También se ha comentado como se pueden aprovechar las propiedades de este medio con los sistemas MIMO. Después se ha elegido la arquitectura utilizada para el receptor y se han definido las propiedades más importantes que deben tener en cuenta Por último se han escogido los componentes ajustándose a las características requeridas y se han realizado las medidas necesarias para comprobar su correcto funcionamiento.

La meta de este proyecto era diseñar un receptor robusto que superase las desventajas que tenían diferentes prototipos diseñados anteriormente debido a propiedades indeseadas como no simetricalidad. Por tanto la simetricalidad ha recibido especial atención a lo largo de todas las fases del diseño.

También se prentedía que el receptor fuera útil ante todas las restricciones que el medio inalámbrico impone. Por eso, para el diseño del receptor se partió de un riguroso estudio de las restricciones de este medio.

Otras de las pretensiones era que el receptor tuviese un interfaz con señales simples para poder trabajar con todos los instrumentos del laboratorio.

Al final de este proyecto todas estas pretensiones han sido cumplidas, es decir, se ha logrado la robustez requerida para el receptor debido a la propiedad de simetricalidad en todos sus componentes, satisface todos los requerimientos impuestos por el medio inalámbrico y es apto para trabajar con todos los instrumentos del laboratorio.

A pesar de que se han logrado muchos de los objetivos, el receptor puede ser optimizado y diferentes puntos han comenzado a ser estudiados para su posible optimización:

la optimización del espacio utilizado es quizás el punto más clave ya que el receptor diseñado ocupa un tamaño considerable y deja ciertos huecos aprovechables para integrar mayor número de componentes

La inclusión de elementos que en este momento son externos como antenas o osciladores locales es otro punto a tener en cuenta en el futuro

Mayor aislamiento entre componentes también puede ser logrado mediante la utilización de ciertas estructuras como el divisor Wilkinson en vez del divisor resistivo. Esto podría mejorar la calidad de la señal, teniendo en cuenta siempre un equilibrio con las pérdidas introducidas por los componentes.

Leibniz Universität Hannover

Institute of Radiofrequency and Microwave Engineering

Master Thesis

Redesign and optimization of a HF receiver for MIMO measurements

Alfonso Rosagro Escámez

Hannover, July 2007

Contents

Chapter 1

Introduction

a. Abstract and motivation

In order to fulfil the steadily growing demand on **transmission bandwidth** and **higher data rates** by means of transmit (TX) and receive (RX) circuitry the Department of High-Frequency-Technology and Radio Systems is working on **MIMO system** (Multiple Input Multiple Output). Herein, transmitter and receiver exploit the **multipath structure**, inherently given by the underlying mobile radio channel, in order to increase the data rate of the communication system or to enhance the RX signal quality (**Diversity**). Especially the group of planar broadband multiarm antennas that have been developed at the Department of High-Frequency-Technology and Radio Systems adapt these transmission schemes because they are capable of providing nearly independent transmission paths in terms of orthogonal antenna arms. The potential of these antenna structures with respect to MIMO communication has already be verified by means of simulation based performance analysis.

As far as this work is concerned, it is desired to verify these simulations by real measurement results in indoor channels. Therefore a basic concept of a MIMO measurement setup (MIMO demonstrator) has been developed and initial work has been started to implement and realize the setup.

The purpose of this work consists of carrying out an indoor wireless MIMO channel sounder. A special RF transceiver will be designed, built up and integrated into the existing laboratory equipment for this purpose. The concept contains a considerable amount of signal processing. Therefore this project has been realized in parallel with another work consisting in signal processing for code synchronization, carrier-recovery as well as frequency and phase estimation and mainly the subsequent estimation of certain MIMO channel parameters.

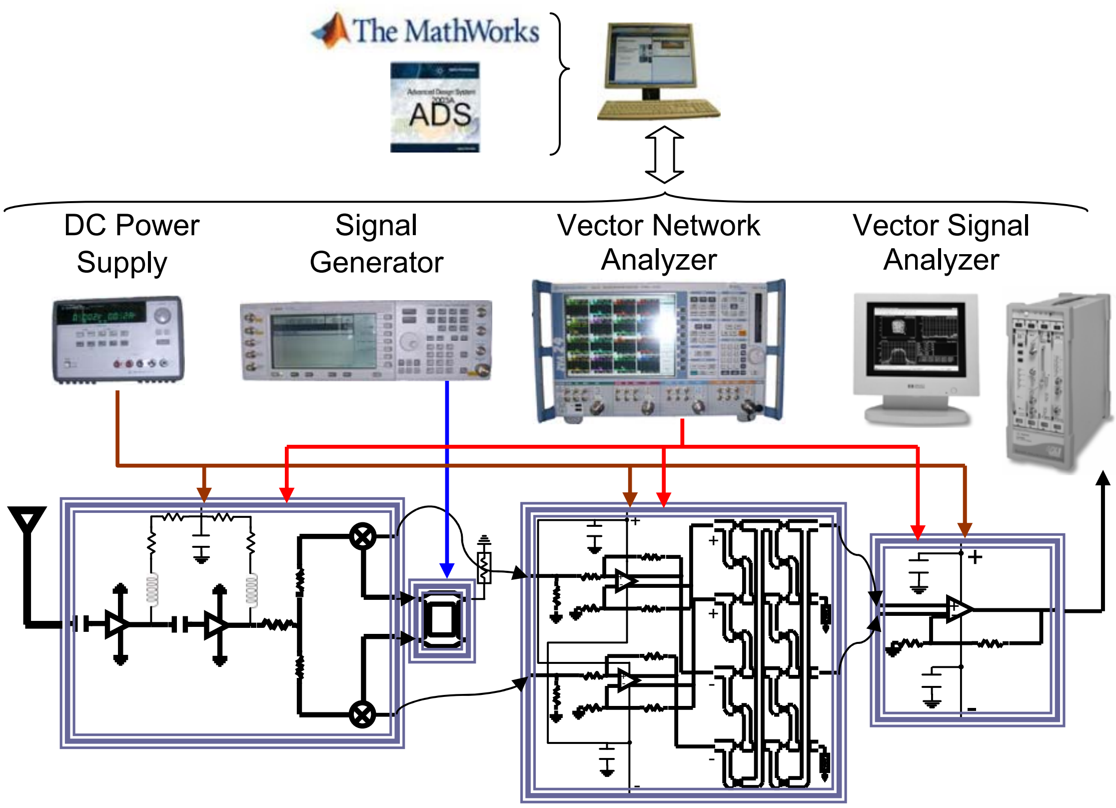


Figure 1.1: Software and Hardware used in the design of the special Receiver

b. Structure

This report describes the design and implementation of a special RF transceiver for MIMO channel sounder, and is organized in 5 chapters. Chapter 2 gives an overview of the wireless mobile radio propagation channel. The chapter describes the most important effects in this channel and the limitations that the wireless channel imposes in the communication. An overview of the MIMO channel is given as well in this chapter. This overview includes the advantages that MIMO offers to exploit the wireless channel and a short explanation of MIMO systems. Chapter 3 discusses the receiver and transmitter architectures considered and points out the strengths and weakness of each one. The motivation behind the choice of a low IF receiver architecture is given in this chapter as well. Chapter 4 is the main chapter of this report since it describes the receiver design. This chapter is divided in two parts. The first part of this chapter covers the theoretical explanation of the desired characteristics of a receiver and the aspects which have influence on them. It includes the modulation scheme used in the transmission, the receiver requirements because of the impairments of the wireless channel and some characteristics of the devices used. The specifications and a detailed planning of the receiver are given in the second part of the chapter.

Chapter 2

The wireless Mobile Radio Propagation Channel

When designing a transceiver, the properties of the mobile communication environment have to be taken into account. A good knowledge of wave propagation and signal transmission over the wireless channel from the transmitter to the receiver and vice versa are of great importance for wireless system design. Therefore in this chapter the main properties in the wireless radio channel are explained.

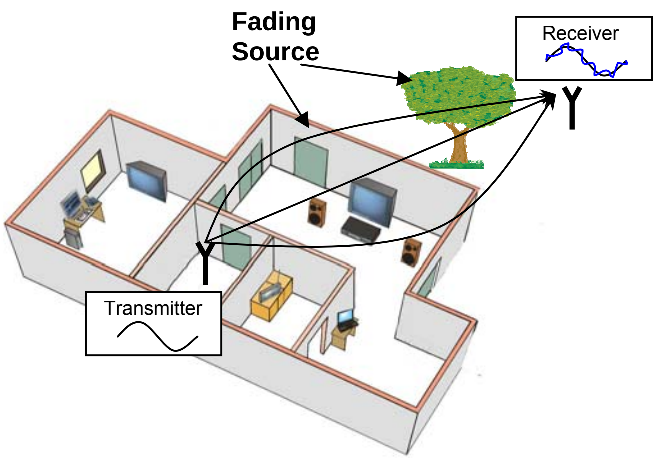


Figure 2.1: Fading sources

a. Fading effects in wireless mobile channels

When a signal is transmitted from transmitter to receiver, it travels over multiple paths, in other words, the signal offered to the receiver contains not only a direct line of sight wave, but also a large number of reflected waves. These reflected waves interfere with the direct wave, which causes different effects, for example random fluctuations in the amplitude, phase and angle of arrival. This phenomenon is named multipath fading.

The two types of fading effects that characterize mobile communication and the wireless channel are large-scale and small-scale fading. In a transmission over large areas, the signals processed experiences both types of fading and it is necessary to take different contributions into account when making a link budget for wireless communication terminals. An example for a receive signal under fading is given in Fig. 2.2:

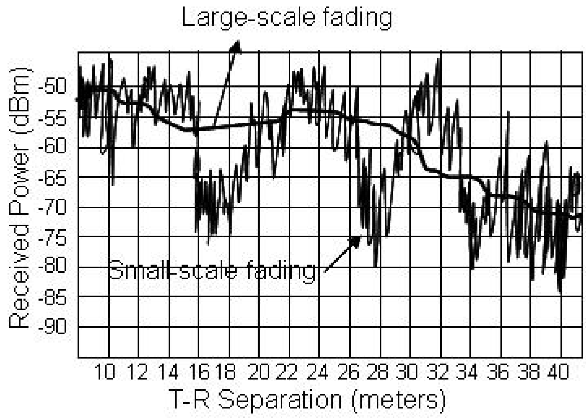


Figure 2.2: Example of received signal power under fading effects.

The contributions which can be appreciated in the picture are:

- Mean path loss as a function of distance due to large scale fading

- Variations about the mean path loss, or large scale fading margin.

- Rayleigh or small scale fading margin.

As can be seen the received signal is affected by two fading mechanism, one of them is the local mean or log-normal fading and the other one is the multipath or Rayleigh fading that results in a representation of the signal where the small scale fading is superimposed on large scale fading. In the next diagram, the probability of the received power in a mobile station is represented:

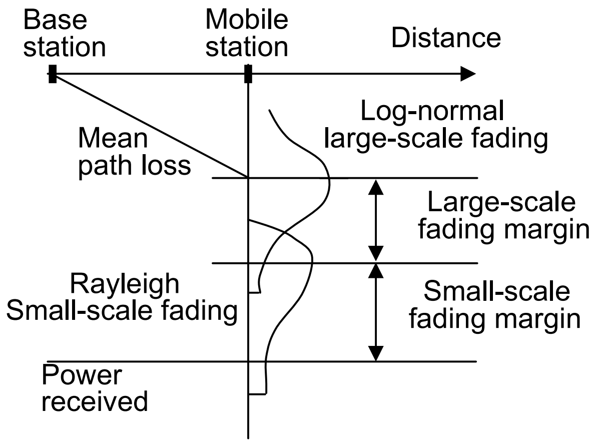


Figure 2.3: Probability of the received power for a determined distance.

The contributions appreciated in Fig. 2.2 can be appreciated in this Fig. 2.3 as well. In this case, the variation about the mean path loss and the small scale fading represented by their probability function can be seen. These probability functions are added to the mean path loss at a determined distance indicating the probability of the power signal at this point.

Therefore, the received power at a determined distance can be seen as a probability function depending on the three components mentioned.

In the next pages a thorough explanation of these effects is done.

Small scale fading represents a kind of fading which occurs within small movements of a mobile or obstacle. This phenomenon provokes dramatic changes in signal amplitude and phase as a result of small changes in the spatial separation between a receiver and transmitter.

To characterize the behavior of the channel caused by the small scale fading a lot of studies have been carried out to find an appropriate statistical characterization. However, the most common statistical description is the Rayleigh probability density function, which describes the envelope of the received signal when multiple reflective and scattered waves exist and there is no line-of-sight component. Thus, for a single link it represents the probability density function (PDF) associated with the worst case of fading per mean received signal power.

The mobile Rayleigh channel is characterized by rapidly changing channel characteristics. If a certain minimum (threshold) signal level is needed for acceptable communication performance, the received signal experiences periods of sufficient signal strength or "non-fade intervals" and insufficient signal strength or "fades". It is of critical importance to the performance of mobile data networks that the used packet duration is selected taking into account the expected duration of fades and non-fade intervals. An example of a received signal with non-fade and fad intervals is given in Fig. 2.4:

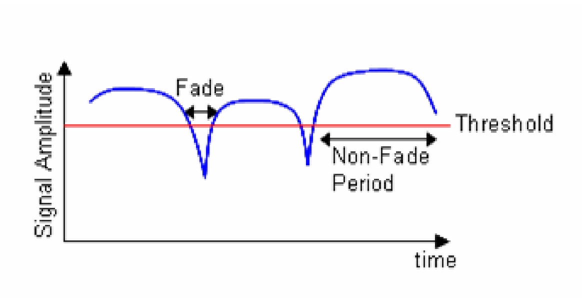


Figure 2.4: Fade and non-fade duration for a sample of a fading signal.

In the next figures can be seen the effect of an antenna displacement when the Rayleigh probability density function is considered to describe the envelope of the received signal:

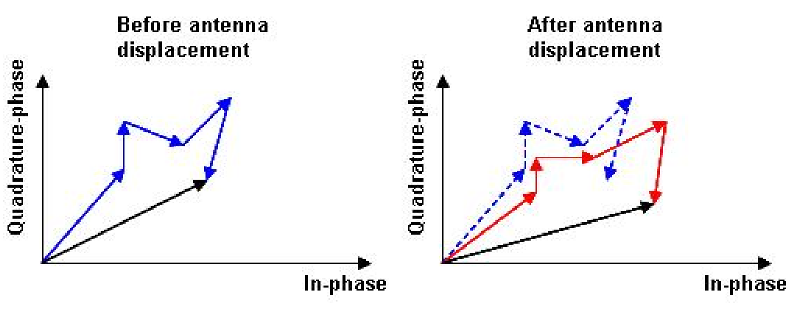


Figure 2.5: Phasor diagram of a set of scattered as result of antenna displacements

The diagram located to the left of Fig. 2.5 shows a phasor diagram of a set of scattered waves before antenna displacement (blue), resulting in a Rayleigh-fading envelope (black). In the diagram located to the right of Fig 2.5 a new set of scattered waves is superposed after motion, resulting in a new Rayleigh-fading envelope.

Two mechanisms characterize the small scale fading itself; **time spreading** of the signal (signal dispersion) of the underlying digital pulses within the signal and **time variant** behavior of the channel.

These two small scale fading mechanisms can be characterized in the two domains (time or time-delay and frequency or Doppler shift). Each mechanism can exhibit different degradation categories.

To understand these phenomena it is necessary to distinguish between two different time references, **delay time** and **transmission or coherence time**. Delay time refers to the time-spreading manifestation which results from the imperfect channel impulse response and imposes a lower limit in the symbol time. The transmission time, however, is related to the time-varying nature of the channel caused by relative motion between a transmitter and receiver, or by movement of objects within the channel and imposes an upper limit in the symbol time. When a narrow pulse is transmitted and the antenna position is modified the delay time in the received pulse as well as the transmission time are different. In this case the response pattern differs significantly in the delay time of the largest signal component, the number of signal copies, their magnitudes and the total received power (area) in the received power profile.

Because of multipath reflections, the channel impulse response of a wireless channel looks like a series of pulses in the time domain. In the next diagram an example of impulse response and frequency transfer function of a multipath channel is shown.

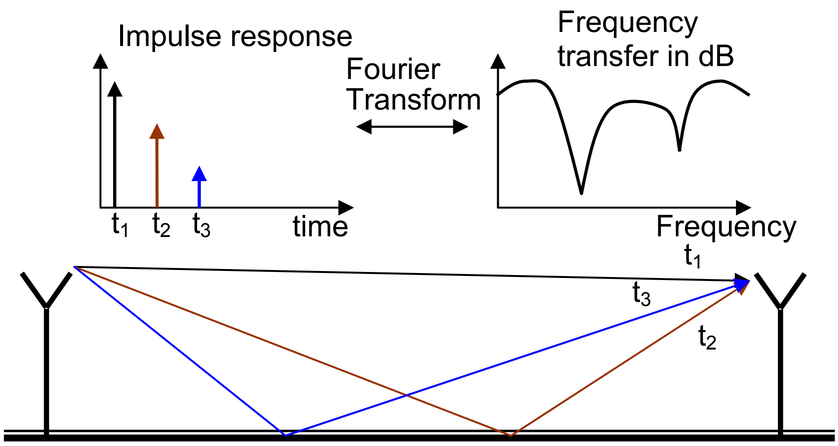


Figure 2.6: Small scale effect: Time spreading

This multipath delay spread characterizes the **time spreading mechanism**. In the frequency domain the channel is characterized as a channel coherence bandwidth (the bandwidth over which the channel transfer function remains virtually constant) as can be observed in the previous figure.

The relationship between maximum delay time (the total time interval during which reflections with significant energy arrive), Tm, and symbol time, Ts, can be viewed in terms of two different degradation categories, frequency-selective fading and frequency nonselective or flat fading.

A channel is said to exhibit frequency-selective fading if Tm > Ts. This condition occurs whenever the received multipath components of a symbol extend beyond the symbol's time duration. Such multipath dispersion of the signal yields the same kind of ISI distortion which is known from electronic filter. In the case of frequency-selective fading, mitigating the distortion is possible because many of the multipath components are resolvable by the receiver.

A channel is said to exhibit frequency nonselective or flat fading if Tm < Ts. In this case, all the received multipath components of a symbol arrive within the symbol time duration. Therefore, the components are not resolvable. Here, there is no channel-induced ISI distortion and the effect in this case is loss in SNR.

A completely analogous characterization of signal dispersion can be made in the frequency domain as mentioned above. In this domain, the coherence bandwidth Bc, is defined as a statistical measure of the range of frequencies over which the channel passes all spectral components with approximately equal attenuation and linear phase. A channel is said to exhibit frequency-selective fading if Bc < 1/Ts where 1/Ts is the symbol rate or signal bandwidth.

When the symbol rate of a system is less than the coherence bandwidth of the channel, the system can be considered to be narrowband. When the symbol rate of a system is more than the coherence bandwidth of the channel, the channel is said to be wideband.

The **time-variance** is characterized as a channel coherence time (T_c), a Doppler spread in frequency domain. This mechanism defines the interval of time in which the channel impulse response remains nearly constant or in other words, for how long a continuous wave can be transmitted without changes in the amplitude and phase being expected. Therefore, the time-varying channel decorrelates after approximately T_c seconds.

If $A_c(\Delta t)$ is an autocorrelation function defining how the channel impulse response decorrelates over time, the channel coherence time (T_c) define the range of values over which $A_c(\Delta t)$ is approximately nonzero. $A_c(\Delta t = T) = 0$ indicates that observations of the channel impulse response at times separated by T are uncorrelated and therefore independent.

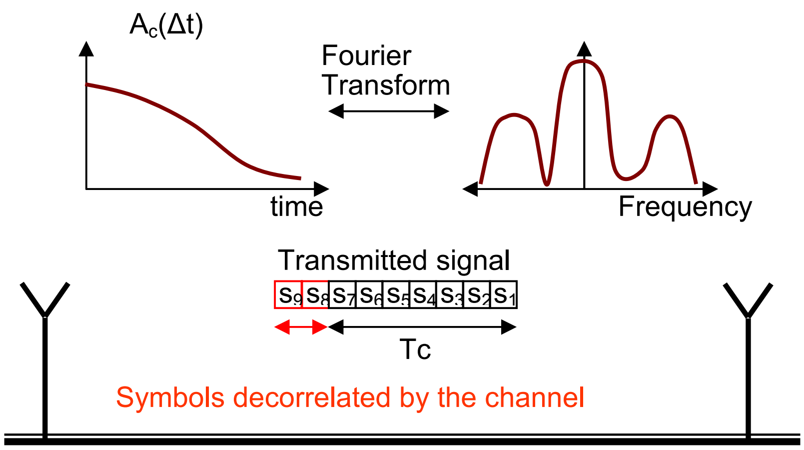


Figure 2.7: Small scale effect: Time variant

A channel is said to exhibit fast fading behavior when Tc<Ts. That is, the time duration in which the channel behaves in a correlated manner is short compared to the time duration of a symbol. It can be expected that the fading character of the channel will change several times while a symbol is propagating due to movement of objects within the channel, leading to distortion of the baseband pulse shape, resulting in a loss of SNR, which introduce an irreducible error rate. Distortion takes place because the received signal components are not all highly correlated throughout time. Such distorted pulses cause synchronization problems in addition to difficulties in adequately defining a matched filter.

A channel is generally referred to as introducing slow fading if $T_c > T_S$. Here, the time duration wherein the channel behaves in a correlated manner is long compared to the time duration of a transmission symbol. Here, it can be expected that the channel state remains constant during the time in which a symbol is transmitted. The propagating symbols are unlikely to suffer from the pulse distortion described above. The primary degradation in a slow-fading channel is loss in SNR.

A completely analogous characterization of the time variant nature of the channel can be made in the frequency domain. The reader is referred to the literature in the bibliography of this chapter to obtain more detailed information about these effects.

Large scale fading is a kind of fading that represents the average signal power attenuation or path loss caused by larger movements of a mobile or obstructions within the propagation environment.

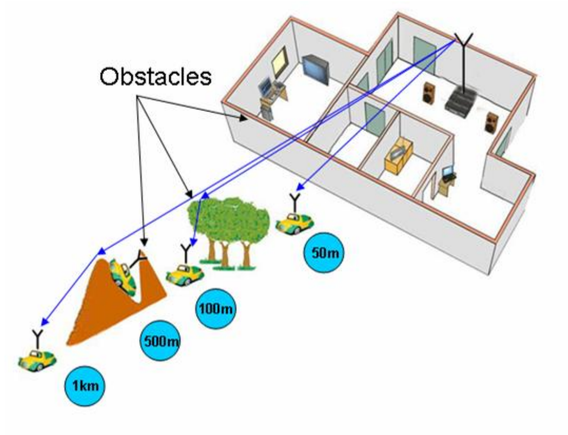


Figure 2.8: Large scale effect

This is often modelled as a mean path loss and a log normally distributed variation. The knowledge of the statistics of large scale fading provides a way of computing an estimate of path loss as a function of distance.

In this case the narrowband statistics can be used to calculate the signal power attenuation for a mobile radio channel. The easiest way to carry out the path loss measurements for a wide range of antenna heights and coverage distances is the Okumura-Hata model.

In this model the average path loss L_p is estimated over a multitude of different sites for a given value of the distance, d, as can be seen in the next picture:

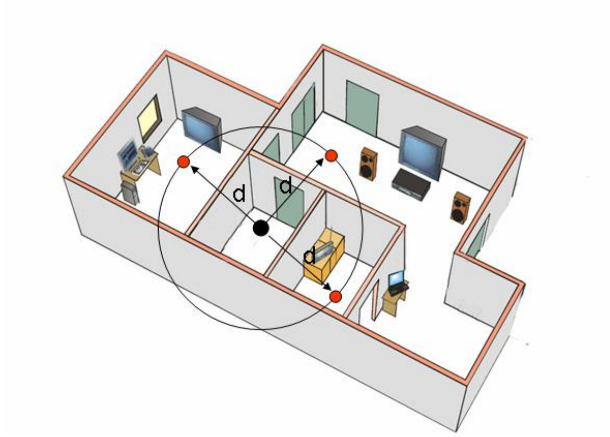


Figure 2.9: Different sites for one determined distance

This model defines the mean path loss L_p as a function of the distance between the transmitter and receiver with the following relationship:

$$\overline{L}_p = \frac{d}{d_0}^n \xrightarrow{DECIBELS} \overline{L}_p(dB) = L_p(d_0)(dB) + 10 \cdot n \cdot \log(d / d_0) \qquad (2.1)$$

In the next diagram the different parameters of the relationship are represented:

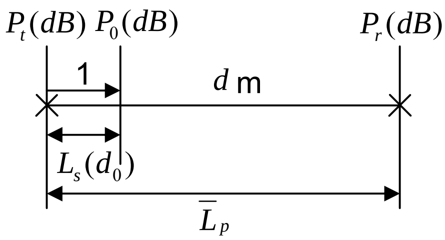


Figure 2.10

The reference distance d_0 corresponds to a point located in the far field of the antenna. Typically, the value of d_0 is taken to be 1 km for large cells, 100 m for microcells, and 1 m for indoor channels (picocells).

Linear regression for a minimum mean-squared estimate (MMSE) fit of $\overline{L}_p(d)$ versus d on a log-log scale (for distances greater than d_0) yields a straight line with a slope equal to 10n dB/decade.

The value of the exponent n depends on frequency, antenna height, and propagation environment. In free space, n = 2 and in the presence of a very strong guided wave phenomenon (like urban streets), n can be lower than 2 and when obstructions are present, n is larger (> 2).

An example of a possible average received power versus distance, by using this model and a typical value of n = 2, is represented in the next figure:

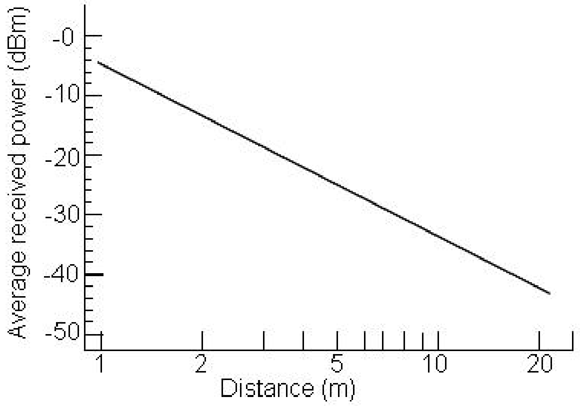


Figure 2.11: Linear regression: $\overline{L}_p(d)$ vs $d(m)$ (log-log scale)

The path loss $L_p(d_0)$ is typically estimated through field measurements or calculated using the free-space path loss from the following expression:

$$L_p = \left(\frac{4\cdot\pi\cdot d}{\lambda}\right)^2 \xrightarrow[DECIBELS]{d=1} L_s(dB) = 20\cdot\log(4\cdot\pi/0.3) + 20\cdot\log(f(GHz) \quad (2.2)$$
$$= 32.44dB + 20\cdot\log(f(GHz)$$

The relationship outlined above provides a mean value, which is not adequate when a concrete signal path is characterized. In this case, to introduce a random variable with a log-normal distribution is necessary. The zero-mean Gaussian random variable Xs with standard deviation σ and distance dependence provides for variation about the mean since the environment of different sites may be quite different for similar transmitter-receiver separations. An example of a log-normal density function can be observed in the next graphic for standard deviation σ = 1 and mean μ = 0:

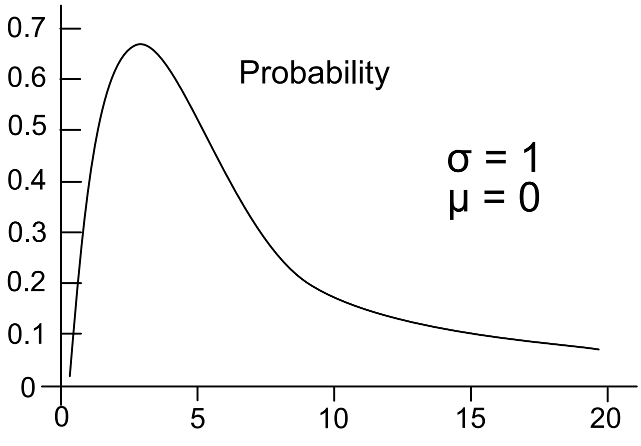


Figure 2.12

Thus, path loss $\overline{L}_p(d)$ can be expressed in terms of $\overline{L}_p(d)$ plus a random variable Xs, as follows:

$$\overline{L}_p = \frac{d}{d_0}^{n} \xrightarrow{DECIBELS} \overline{L}_p(dB) = L_s(d_0)(dB) + 10 \cdot n \cdot \log(d / d_0) + X_s \qquad (2.3)$$

Based on measurements values as high as 6-10 dB or greater can be observed for Xs.

b. MIMO System

The communication channel is an important component in order to determine a good system performance. It is possible in wireless communication to achieve better performance. A MIMO (Multiple Input – Multiple Output) system is a practical multi-antenna link to exploit the channel space-time resources but including a higher level of complexity in the channel space-time characteristics.

The communication behaviour of a MIMO channel includes the properties of the propagation environment, the impact of antenna properties, the antenna array configuration and the radio frequency architecture.

A generic MIMO communication system can be seen in the following diagram:

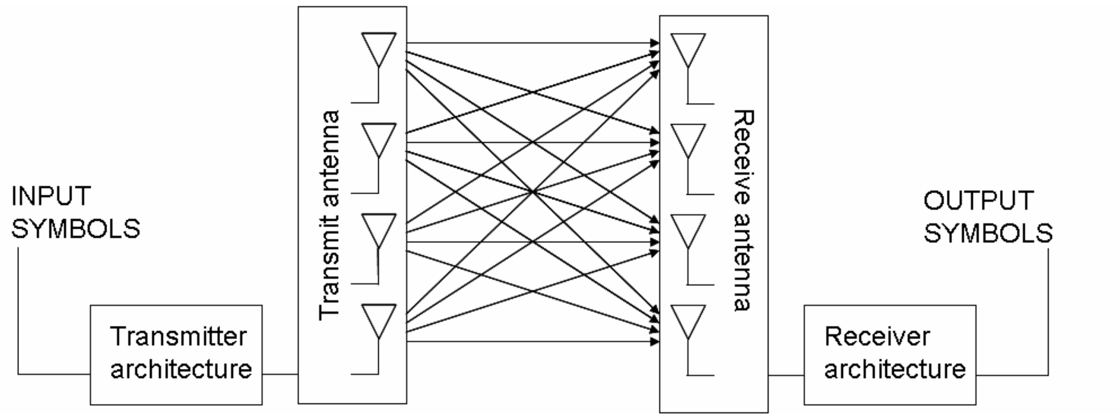


Figure 2.13: Generic MIMO communication system

In this system a stream of vector input time dependent symbols, are fed into a space-time encoder, generating a stream of NT × 1 complex vectors, where NT represents the number of transmit antennas. Each element of the vector is transformed with pulse shaping filters to create a NT × 1 time-domain signal vector, which is up-converted to a suitable carrier frequency. These signals are lead to the antennas, which radiates energy into the propagation environment.

At this moment the radiated field is affected by the effects explained above as well as the elevation and the azimuthal angles taken with respect to the coordinate frame of the transmit array (θT, δT) and the receive array (θR, δR). To know the value of the field incident distribution on the receive array is possible by means of the convolution between the field radiated and the impulse response of the channel, which is a function of the variables outlined above.

The NR-element receive array then samples this field and generates the NR × 1 signal vector at the array terminals. Noise in the system is typically generated in the physical propagation channel (interference) and the receiver front-end electronics (thermal noise). To simplify the explanation all additive noise can be lumped into a single contribution represented by the NR × 1 vector η(t) that is injected at the receive antenna terminals. The resulting signal plus noise vector is then down converted and post processed to produce the NR × 1 output vector, which is demodulate and decoded to estimate the originally transmitted symbols.

The best way to understand the MIMO wireless channel is to experimentally measure a NR x NT system. When measuring the channel properties the effects of the RF subsystem and the antennas are included in the results obtained and a variation in the array configurations used has an effect in these results. By this way, different measurements of channel capacity, signal correlation structure (in space, frequency and time), channel matrix rank, path loss, delay spread, and a variety of other quantities can be carried out. With these measurements a model of the MIMO channel can be implemented. There are several tactics to carry out the measurements and each one has its advantages and disadvantages.

The true array system consists in measuring when all antennas operate simultaneously, and is the most closely model real world MIMO communication. Its main advantage is that it can accommodate channels that vary in time, however, the implementation of such a system comes at significant cost and complexity because of the requirement of multiple parallel transmit and receive electronic subsystems. Other techniques that can be used in order to develop a model for a MIMO channel can be either “switched array” or “virtual array” architectures. “Switched arrays” use a single transmitter and single receiver to carry out the measurements and sequentially connecting all array elements to the electronics using switches. Switching times must be low to carry out the measurements over all antenna pairs while the channel remains constant. “Virtual arrays” use precision displacement (or rotation) of a single antenna. Although Its main advantage is the non-existence of mutual coupling, this architecture usually take several second or even minutes and therefore, is only appropriate for fixed indoor measurement campaigns when the channel remains constant for a long time (long coherence time).

The channel performance is determined not only for the channel but also for the antenna system. When using two antennas, the received signals are given as:

$$s_1 = \sum_{l=1}^{L} E_l \left[e_1(\theta_l, \phi_l) \cdot \bar{e}_l \right] e^{-j(\pi d/\lambda)\sin\theta_l \cos\phi l} \tag{2.4}$$

$$s_2 = \sum_{l=1}^{L} E_l \left[e_2(\theta_l, \phi_l) \cdot \bar{e}_l \right] e^{j(\pi d/\lambda)\sin\theta_l \cos\phi l} \tag{2.5}$$

Where $e_1(\theta_l, \phi_l)$ and $e_2(\theta_l, \phi_l)$ are the vector field patterns of the antennas, and (−d/2, 0, 0) and (d/2, 0, 0) are the coordinates where the antennas are placed. L is the number of waves which impinge on the antenna array and E, θ1, φ1 and ê are the complex field strength, the arrival angles and the electric field polarization of the waves respectively.

The received signal s1 and s2 must be unique despite the fact that both antennas observe the same set of plane waves, in order to get an effectively work for the MIMO system.

As can be seen in the expressions above, it is possible to play with the antennas to accomplish this goal providing a unique weighting to each of the plan waves. Based on the expression above, there are three different ways to do it. **Spatial diversity**, consisting in different element position, each antenna places unique phase on each multipath component based on its arrival angles, **pattern diversity**, consisting in the same polarization for both antennas but different magnitude and phase responses (in different directions ("angle diversity")) and the last mention is **polarization diversity**. Here the antenna pattern differs only in their polarization for separating the different signals.

In this work, the true array system is used to carry out a modular build up of a two channel transceiver for wireless indoor channel sounding, MIMO and Diversity application in the 2.45GHz ISM-Band. In the chapter dedicated to channel sounding measurements a thorough explanation for the measurement technique of a MIMO channel is given. Once the IF data are collected, post processing is used to perform carrier and symbol timing recovery and code synchronization. The explanation of the postprocessing algorithms is out of the objectives of this work. The algorithms used are explained in the reference "Signal processing for the estimation of the narrowband MIMO channel" where the program Matlab is used for this purpose.

Through these measurements, a channel model, which capture the key behavior observed in the experimental data, can be done and this allows performance assessment of potential space-time coding approaches in realistic propagation environments. There are several kinds of models that can be used to characterize the MIMO wireless channel depending on the tradeoffs required. For example models such as Random matrix, geometric discrete scattering or statistical cluster models. In order to obtain a bigger knowledge about these models, turn to the references.

Chapter 3

Transceiver architectures

Several different architectures can be used when designing a RF transceiver. The choice of the architecture affects the complete system. Therefore this step has very high importance. Every architecture has its own advantages and disadvantages, and the choice depends on the application. Several factors have to be considered, such as power consumption, performance, cost, size, weight, integration level, complexity and time to market. The relative weight of all other factors is determined by the application at hand.

In the following paragraphs, before the proposed design is chosen, the different possibilities of receiver architectures are explained. After that, the proposed design will be explained briefly.

When a modular build up of a channel transceiver is designed, the receiver architecture has the most stringent requirements. Therefore, in the next chapter, a thorough explanation about the different possibilities of receiver architectures is done.

3.1 Receiver architecture

a. Homodyne architecture or direct conversion

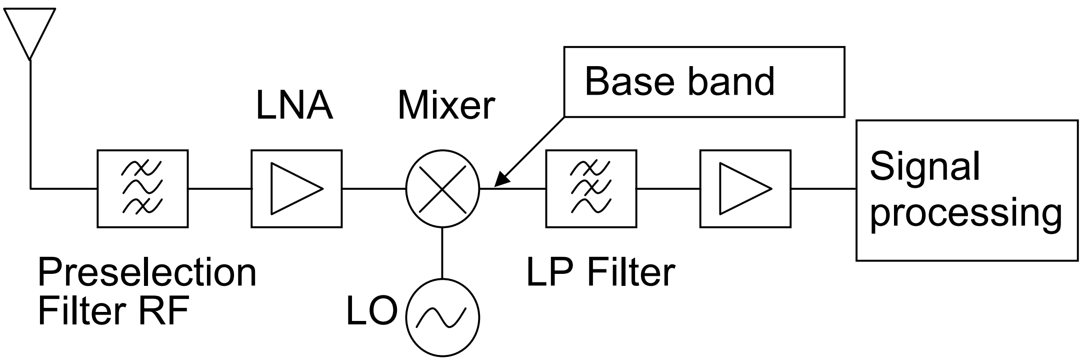


Figure 3.1: Block diagram of a homodyne architecture or direct conversion

The direct conversion architecture is the simplest way to design a receiver. The mode of operation of this architecture consist of down-converting the signal directly to base-band by mixing the received RF signal with a local oscillator at the same frequency and low-pass filter the achieved base-band signal. Usually, the mixers have a very high noise. Therefore they are preceded by a Low Noise Amplifier (LNA). Although by doing direct conversion there is no image frequency in a strict sense, the lower half of the signal spectrum itself is folded to the positive frequency axis and overlaps with the upper half of the spectrum. In order to distinguish the lower from the upper half of the spectrum and thus recover the information provided by the phase of the signal, a vector detection scheme with quadrature LO signals and in-phase (I) and quadrature (Q) baseband paths is required.

When frequency and phase information is required, the RF signal is split up into 2 different paths and mixed in one path with the LO and in the other path with its quadrature, so that the signal is received in the output in phase and quadrature. This is necessary in phase and frequency modulation schemes.

Although this architecture is the simplest one and it is possible to obtain the highest integration, the use of this architecture has some drawbacks that require particular consideration. Its main drawbacks are explained in the next paragraphs:

DC Offsets.

Since in a homodyne receiver the signal is down-converted directly to base-band (in the vicinity of the zero frequency), different offset voltages can corrupt the signal and saturate the following stages. To understand the origin and impact of offsets, it is necessary to know the effect called “LO leakage”. This effect can be seen in the next figure:

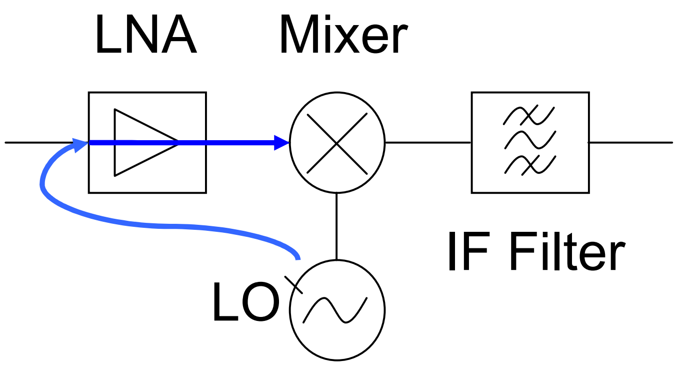


Figure 3.2: LO leakage

“LO leakage” is the term for the effect, which occurs when a finite amount of feedthrough exists from the LO port to other points in the circuit. In other words, the isolation between the LO and RF ports of the mixer is not perfect. This effect takes place due to capacitive coupling. The leakage signal appearing at the input of the LNA is amplified and mixed with the LO signal, thus producing a DC component at the output of the mixer. This phenomenon is called “self-mixing”. This amount of DC is quite high due to the gain in the whole receiver. The problem becomes very important because in most of today’s modulation schemes the spectrum contains information at frequencies as low as a few tens of hertz, mandating a very low corner frequency in the filter. In addition to difficulties in implementing such a filter a more fundamental problem is its slow response, an important issue if the offset varies quickly. For these reasons, homodyne receivers require sophisticated offset-cancellation techniques. Modulation schemes with negligible energy below a few kilohertz can prevent this problem, by allowing the use of AC Coupling.

LO re-radiation (also afflicting low-IF receivers).

This effect can be seen in the next figure:

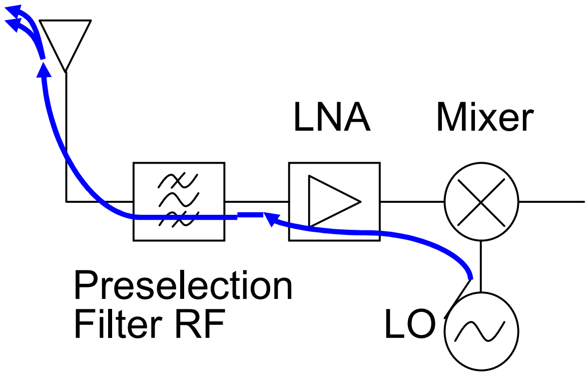


Figure 3.3: LO re-radiation

Because of the limited reverse isolation of mixers and LNAs, some fraction of the LO power may leak to the antenna. The re-radiated signal lies at the center of the channel and may thus disturb nearby receivers tuned on the same frequency.

Even order distortion.

While third order mixing due to the nonlinearity exhibited from the LNA is considered as a source of interference in every architecture, even-order distortion also becomes problematic in homodyne receivers. If two strong interferers close to the channel of interest experience a nonlinearity such as $y(t) = a_1\, x(t) + a_2\, x^2(t)$, then they are translated to a low frequency before the mixing operation and the result passes through the mixer with finite attenuation. This is because, in the presence of mismatches that degrade the symmetry of the mixer, the mixing operation can be viewed as $x(t) \cdot (a + A \cdot \cos(a \cdot t))$, indicating that a fraction of x(t) appears at the output without frequency translation. Another issue is that the second harmonic of the input signal (due to the square term in the above equation) is mixed with the second harmonic of the LO output, thereby appearing in the baseband and interfering with the actual signal.

I-Q Mismatch.

In the phase and frequency modulation schemes, a quadrature mixing is necessary. For that, I and Q phases of the LO are necessary and this implicates the possibility of mismatches between both signals. If the amplitudes of the I and Q outputs are not equal or their phase difference deviates from ninety degrees, the error rate in detecting the baseband signal rises.

Flicker Noise.

Device flicker noise becomes significant at low frequencies. Therefore amplification of the baseband signal with low noise is an important issue.

b. Heterodyne architectures

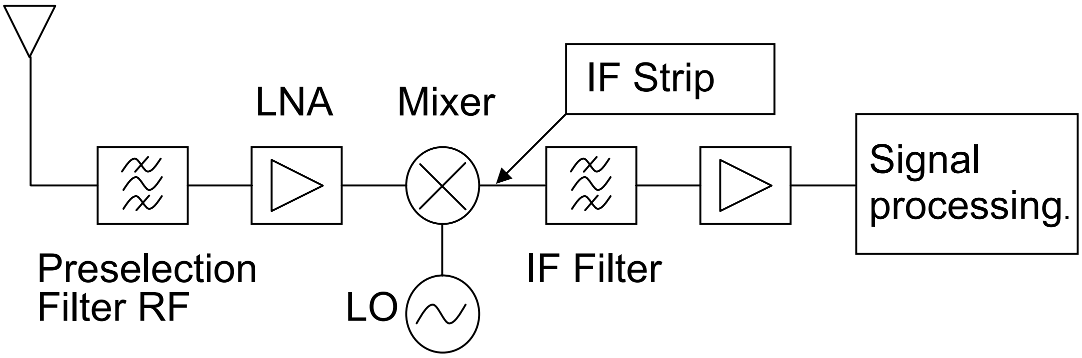


Figure 3.4: Block diagram of a heterodyne architecture

The heterodyne architecture is a more complicated architecture than the direct conversion architecture, but it avoids the drawbacks outlined above. The mode of operation of this architecture is to down-convert the RF signal to an intermediate frequency. After that, the signal is band-pass filtered and amplified and down converted another time to baseband. In the case of frequency or phase modulation schemes, the last down-conversion have to generate both I and Q phases of the signal, causing the problems explained above.

This architecture allows the DC offset in the first few stages to be removed by band pass filtering at the intermediate frequency and the DC offset of the last stage is suppressed by the total gain in the preceding stages.

The I-Q mismatch is also alleviate because this occurs at much lower frequencies and is therefore easier to control and correct.

The problem of the LO re-radiation is alleviated too. Since the frequency of the LO is out of the band of interest, it is suppressed by the front-end BPF and its radiation from the antenna is less problematic.

The best feature of this architecture is its selectivity. When a direct conversion is used, the use of a filter to eliminate a possible high interferer near to the band of interest becomes impossible due to the necessary high quality factor of the required filter. The heterodyne architecture allows the elimination of this interference at a lower frequency, where filters with high selectivity can be used with much more relaxed requirements.

The receiver is tuned by changing the frequency of the LO. Due to the fact that the main selectivity is provided by the channel filter at the fixed IF, no tuning is needed in the channel filter. This is a significant advantage, since the implementation of a filter with a fixed passband characteristic is much easier than the implementation of a tuneable filter.

Some drawbacks are introduced for this architecture as well. The first problem is the image-rejection. A detailed explanation about the image rejection process used in this work is given in a later chapter.

When the received signal is down-converted, both the signal at the frequency of the local oscillator plus the intermediate frequency (wished signal (f_{LO} + f_{IF})) and minus the intermediate frequency (image signal(f_{LO} - f_{IF})) are down-converted to the intermediate frequency because of the mirror properties of the spectrum. As at the output of the mixer the two signals can no longer be discerned, the unwanted signal, called image, has to be suppressed prior to mixing. The preselection filter in this case has to suppress strong out-of-band signals which could saturate the front-end of the receiver and the image signal, located at a frequency of $2f_{IF}$ apart from the desired one. This effect can be seen in the next figure:

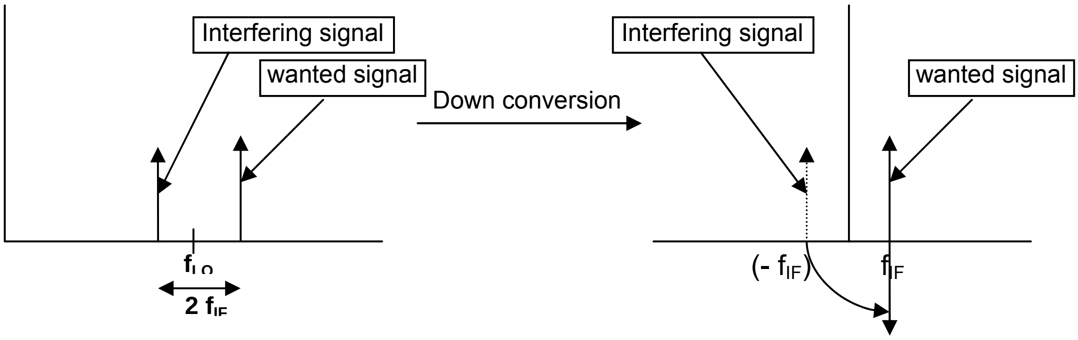


Figure 3.5: Problem of the image frequency in the down conversion

Another drawback of the heterodyne architecture is that the LNA must be terminated with 50Ω impedance because the image-reject filter cannot be integrated and is therefore placed off-chip. This adds another dimension to the trade-offs among noise, linearity, gain, and power dissipation of the amplifier, further complicating the design. The image-reject and channel-select filters are typically expensive and bulky, making the heterodyne approach less attractive for small, low-cost wireless terminals.

The primary drawback of the heterodyne architecture is its inadequacy for high levels of integration. The required high-Q, low-loss, low-distortion filters are well beyond the capabilities of current integrated technologies. Consequently, external passive high-Q filters are generally used, usually implemented with ceramic filters, SAW filters, or sometimes with LC filters. In order to accommodate standards with different bandwidths, an array of selectable channel filters is required.

Due to the very high performance that this architecture can achieve, it is by far the most widely used receiver architecture since the thirties.

The IF frequency has to be chosen with care because of the image rejection. The issue of image rejection leads to a trade-off among three parameters. The amount of image noise, the spacing between the band and the image (=2 IF) and the loss of the filter. A high IF relaxes the design of the image-reject filter, minimizing the image noise (the filter provides more attenuation at the image frequency) but imposes more stringent requirement on the IF strip. It is also possible to tolerate greater loss in the filter while increasing its Q. On the contrary, a low IF frequency simplifies the design of the IF strip, for example, simplifies the design of the amplifiers, reducing the risk of oscillations and lower power consumption, but makes suppression of the image a difficult task.

When a very low frequency is selected, another architecture must be chosen called lo IF receiver.

c. Low IF architecture

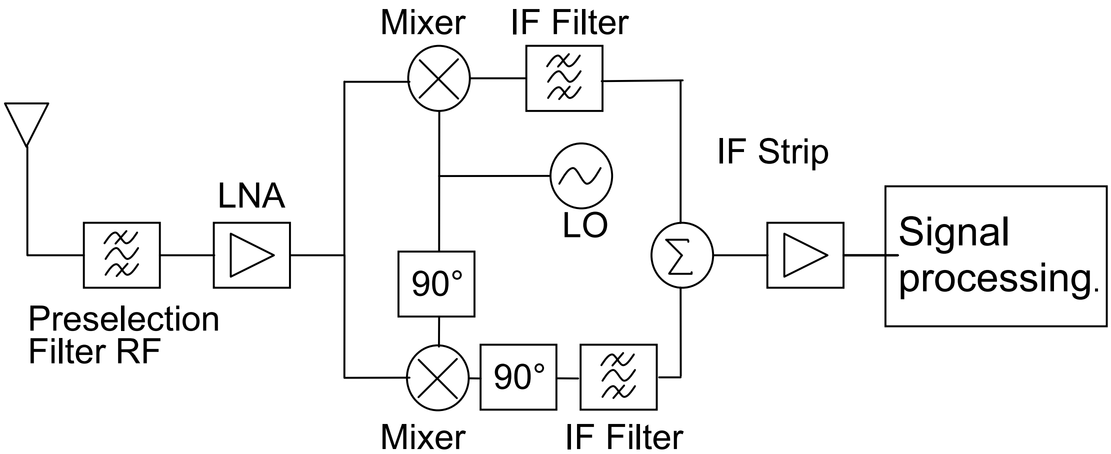


Figure 3.6: Block diagram of a low-IF architecture receiver (Hartley architecture).

When using a very low frequency, the image rejection becomes a critical point. The image frequency is too near to the wanted signal. Therefore, a preselection filter can't be used to reject the image and it is necessary to deal with the image signal in a different way.

There are several techniques available to deal with the image signal. One such technique originates from a single-sideband modulator introduced by Ralph Hartley. The circuit mixes the RF input with the quadrature outputs of the local oscillator, low-pass filters the resulting signals, and shifts one by ninety degrees before adding them. While the components of the wanted signal are added in phase, with ideal I and Q paths, the image is completely suppressed. As the image is suppressed by means of a subtraction, the amount by which it is attenuated is a sensitive function of the gain and the phase mismatch between I and Q paths (including any quadrature error of the LO signals).

If the input is equal to $A_{RF}\cos(\omega_{RF}t)+A_I\cos(\omega_I t)$, where ω_I is the image frequency, the output is proportional to $A_{RF}\cos((\omega_{LO}-\omega_{RF})t)$ (it can be proved mathematically). For typical image reject architecture, the image is rejected by about 30 to 40 dB.

The image rejection ratio (IRR) can be calculated with the help of the following equation:

$$IRR=\sqrt{\sin^2\left(\frac{\delta\varphi}{2}\right)+\left(\frac{\delta G}{2G}\right)^2\cdot\cos^2\left(\frac{\delta\varphi}{2}\right)} \qquad (3.1)$$

Where the δG and δφ are the gain and phase mismatch between the I and Q paths (including any quadrature error of the LO signals).

Another technique was introduced by Weaver as a way for treating the image rejection problem originated when the down-conversion to intermediate frequency is realized. The Weaver technique down-converts the signal in two steps. In the first step, the input is mixed with the quadrature phases of the first local oscillator and the result is coupled with a capacitor to eliminate DC offset and then low-pass filtered. In the second step, these signals are translated to baseband and added together, thereby effecting image cancellation. The important advantage of the Weaver architecture is that it does not require high-Q bandpass filters. This architecture proposed by Weaver can be seen in the next illustration.

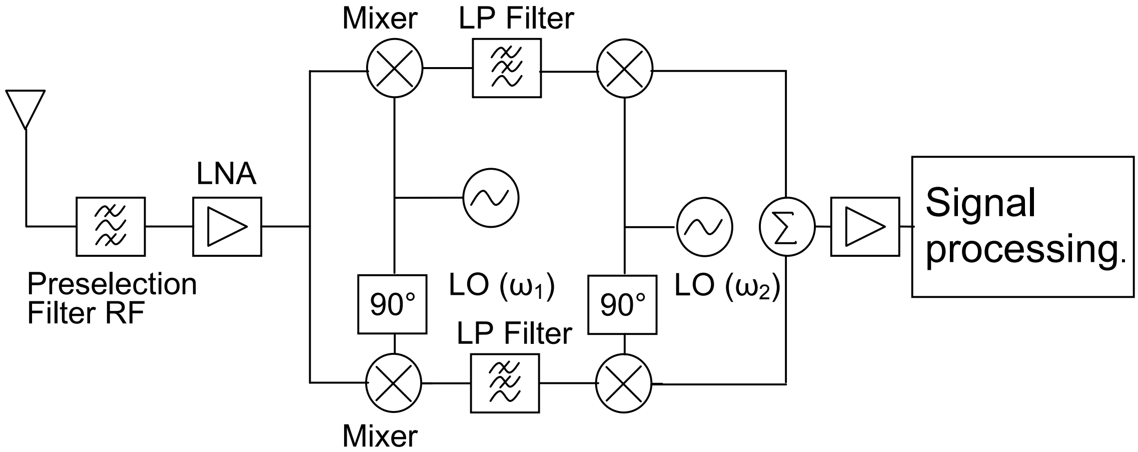


Figure 3.7: Block diagram of a Weaver receiver

Another technique for image reject filtering consists in using a polyphase filter. A thorough explanation of the image rejection using polyphase filter is given in the next chapters.

d. Proposed design

Once the different architectures are explained in general terms, the choice of a concrete architecture becomes a trade-off between its different parameters and the requirements for the RF receiver for its future application.

For the purpose of this work, the most important parameter taken into account is the symmetry of the complete receiver to avoid degradation of the signal which can impede its correct demodulation. In order to simplify the design of the IF strip, a design has been built up with the lowest as possible intermediate frequency. A polyphase filter and several amplifiers have been used to reject the image frequency with a high image reject ratio and to achieve enough gain in the complete system to capture low power signals.

The goal of this chapter is to do a short overview of the receiver designed and the components chosen in order to reach the requirements described above. A thorough explanation of the receiver is in the chapter dedicated to the receiver planning.

The architecture adopted in this work can be seen in the next diagram:

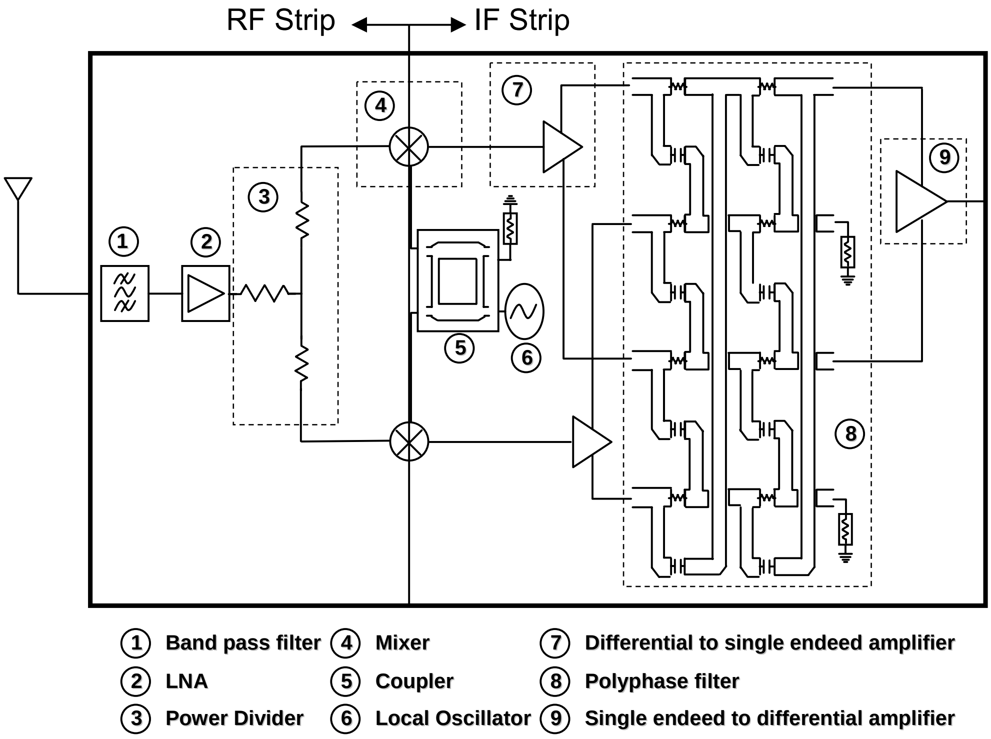


Figure 3.8: Complete receiver chain

As can be seen, in order to avoid the problems described in the direct conversion, a low IF architecture is adopted using a polyphase filter to reject the image.

The components used in the receiver designed are described in the next paragraph:

In the first stage, a band pass filter (1) and a low noise amplifier (2) are placed to suppress the out-of-band frequency components and to get a good signal level when applying the minimum possible of noise in the circuit. In a next step, a resistor power divider (3) is used to divide the output of the amplifier into equal signals, obtaining two different paths (I and Q paths). In every path the signal is down-converted to an intermediate frequency using a mixer (4) and a local oscillator (6). The signal applied to the local oscillator input of the mixer in each path is the in-phase and quadrature component of the local oscillator respectively. To achieve the phase shift required in the local oscillator, a coupler (5) with transmission lines is used. In order to reject the image a polyphase filter (8) is used. The polyphase filter needs the differential signal of both paths. For this a single-to-differential-ended amplifier (9) is used in both paths. At the output of the polyphase filter, a differential signal is achieved. To obtain a single output a differential-to-single-ended amplifier (7) is used.

3.2 Transmitter architecture

Concerning the transmitter architecture, the requirements are much more relaxed than the requirements for the receiver due to the signal is produced in the transmitter and is therefore sufficiently strong. Issues such as image rejection, DC-offset, flicker noise and band selectivity are not so critical, although, the DC-offset needs to be minimized as it gives rise to LO feedthrough which overlaps with the modulated carrier. The critical stage for this architecture is the output power amplifier design. A short explanation about the possible transmitter architectures is done in the next lines.

a. Direct conversion

The direct conversion transmitter architecture is the simplest one and can be used to generate any kind of modulation and is suitable for high integration. It is widely used in mid-high performance transmitters.

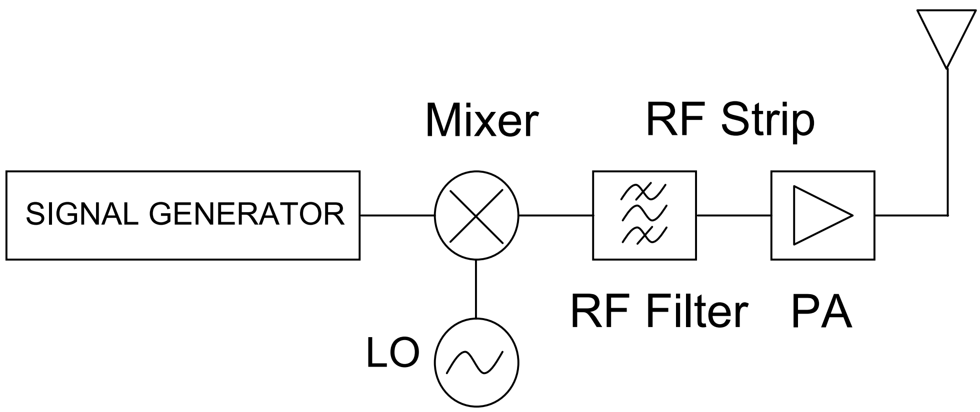


Figure 3.9: Block diagram of a direct conversion architecture

The mode of operation of this architecture consists of mixing the baseband signal directly with the local oscillator output at RF frequency. After that, the signal is band-pass filtered and applied to a power amplifier. The signal in the output of the amplifier is lead to the antenna through a matching network which is interposed between the power amplifier and the antenna to allow maximum power transfer and filter out of band components.

This architecture has some drawbacks. The most important one is the disturbance of the transmitted local oscillator by the output PA. Due to the fact that the output of the PA has quite high power and has a spectrum centered around the frequency of the local oscillator, this signal corrupts the oscillator spectrum. This effect is called “injection pulling” or “injection locking” and is very important due to the mixer tends to shift his operating frequency when an interferer occurs close to the LO frequency.

Different techniques can be used to avoid this problem. The next diagrams show two techniques:

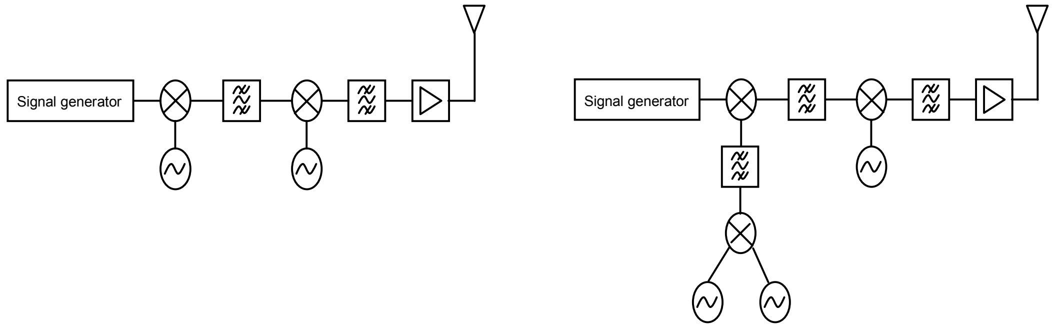


Figure 3.10: Block diagram of two architectures to avoid disturbances of the LO

The diagram located to the right of Fig. 10 is the "two steps conversion architecture". By this way, the base band signal is up converted to the wanted mixing frequency in two steps by using an up-conversion stage with an intermediate frequency. By this way, the frequency in the output of the power amplifier is much higher than the frequency of the two local oscillators, avoiding the effect produced in the mixer when a close interferer is present.

The diagram located to the left of Fig. 10 is another technique to avoid the disturbance of the local oscillator by using the offset-LO principle. This required carrier frequency f_{RF} is generated by two oscillators working at a frequency of $f_{RF} \pm f_{IF}$ and f_{IF} respectively. The desired frequency is then generated with the help of a mixer, while the image needs to be suppressed by a filter.

Different architectures such as open loop modulation or two point modulation can be used to improve the performance of the transmitter but their explanations are out of the purposes this report. For more information about this topic, the reader is referred to literature in the bibliography.

Chapter 4

Receiver design and optimization

When planning to design a RF receiver, an analysis of the relevant requirements of the RF receiver must be done. Once the minimum requirements are fixed, a plan for the receiver can be done. By using the opportunities that the minimum requirements allow, a definitive plan can be done to optimize the important parameters of the receiver for desired application and to assign the required performance figures to each block of the system.

The receiver planning also depends on the modulation scheme used. Therefore a short introduction about the modulation scheme used and its properties is done in the next section. For more information about the modulation schemes, the reference "Signal processing for estimation of narrowband MIMO transmission channel" can be consulted.

After this introduction, an explanation and calculation of the minimum requirements mentioned above are done and the receiver specifications are calculated. The receiver specification also depends on the measurement device collecting the received data. Therefore this chapter includes a section highlighting the vector signal analyzer features used for data acquisition in this work.

At the end of this section a thorough receiver planning is done. The receiver planning includes the components used and its performance, emphasizing in the most critical component of the receiver, the polyphase filter.

a. Modulation scheme

The modulation scheme used in this work is Direct Sequence Spread Spectrum (DSSS) that is the most widely recognized form of spread spectrum. Therefore, in the next paragraphs an overview of the spread spectrum technique is given

A spread spectrum communication block diagram is shown in Fig. 4.1.

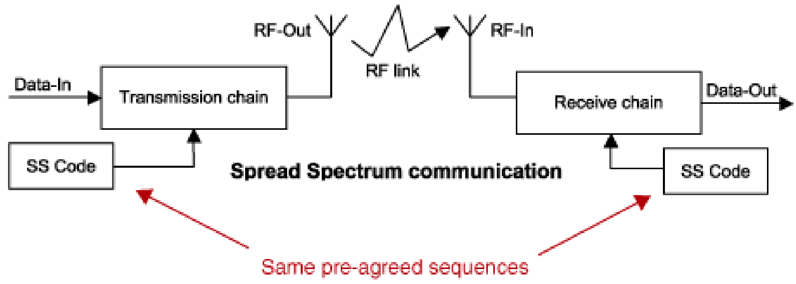


Figure 4.1: spread spectrum communication block diagram is shown below

Spread spectrum (SS) is a form of wireless communications, where the frequency or the phase of the transmitted signal is deliberately varied. These results in a much greater bandwidth than the signal would have occupied if its frequency would not have been varied. The next diagrams show the bandwidth effects of the "spreading" operation:

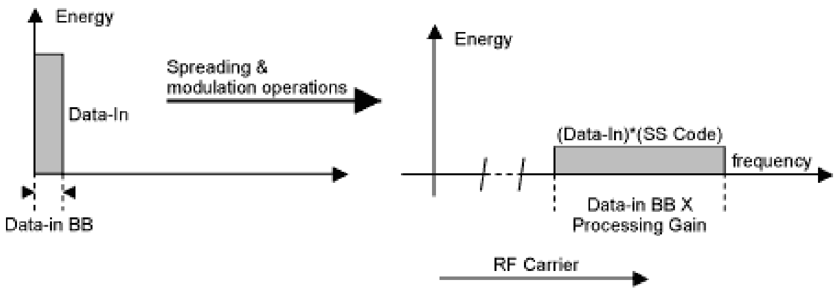


Figure 4.2: bandwidth effects of the "spreading" operation

SS techniques are methods by which energy generated at one or more discrete frequencies is deliberately spread or distributed in time or frequency domains. This is done for a variety of reasons, including the establishment of secure communications, increasing resistance to natural interference and to prevent detection.

In this kind of technique a pseudo-noise code, independent of the information data, is employed as a modulation waveform to "spread" the signal energy over a bandwidth much greater than the signal information bandwidth. At the receiver the signal is "despread" using a synchronized replica of the pseudo-noise code.

The next diagrams show the bandwidth effects of the "despreading" Operation.

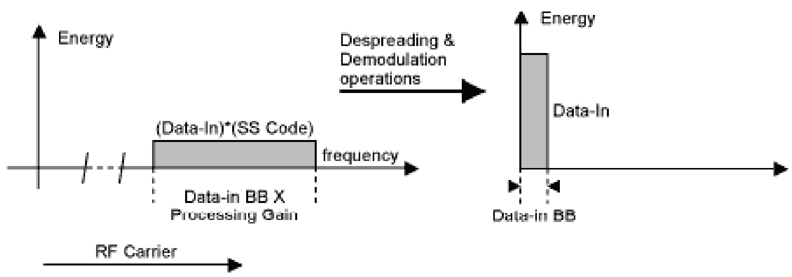


Figure 4.3: Bandwidth effects of the "despreading" operation

All SS techniques have one idea in common, the key (also called code or sequence) attached to the communication channel. The manner of inserting this code defines precisely the SS technique in question. The term "spread spectrum" refers to the expansion of signal bandwidth, by several orders of magnitude in some cases, which occurs when a key is attached to the communication channel. Intentional or un-intentional interference signals are rejected during the despreading operation because they do not contain the SS key. An example of this rejection is shown in the following.

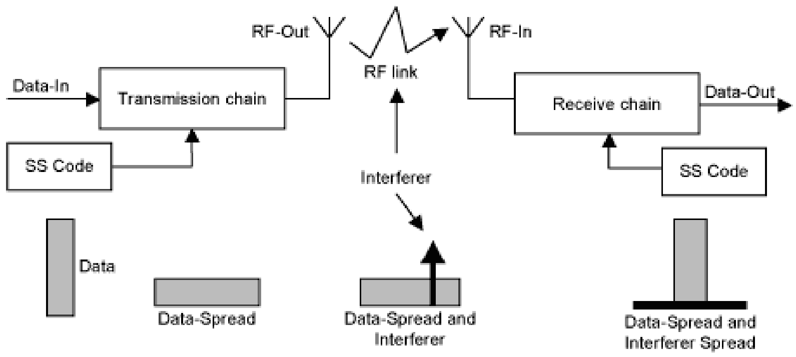


Figure 4.4: example of image rejection

More advantages of these techniques are explained in the reference mentioned above.

As mentioned in the introduction of this section, the modulation technique used in this work is a Direct Sequence Spread Spectrum (DSSS). The DSSS process is performed by effectively multiplying an RF carrier and a pseudo-noise (PN) digital signal. First the PN code is modulated onto the information signal using the technique of modulation BPSK: Then, a mixer is used to multiply the RF carrier and PN modulated information signal. This process causes the RF signal to be replaced with a wide bandwidth signal with the spectral equivalent of a noise signal. The demodulation process (for the BPSK case) is then simply the mixing/multiplying of the same PN modulated carrier with the incoming RF signal. The output is a signal that has a maximum when the two signals are exactly equal one another or are “correlated”. The correlated signal is then filtered and sent to a BPSK demodulator.

The choice of the BPSK modulation is due to the robustness of this modulation. Higher data-rates can be achieved with other techniques, but the robustness of this kind of modulations is much lower.

The spectral content of an SS signal is shown in Fig 4.5. Note that this is just the spectrum of a BPSK signal with a $\left(\frac{\sin x}{x}\right)^2$ form.

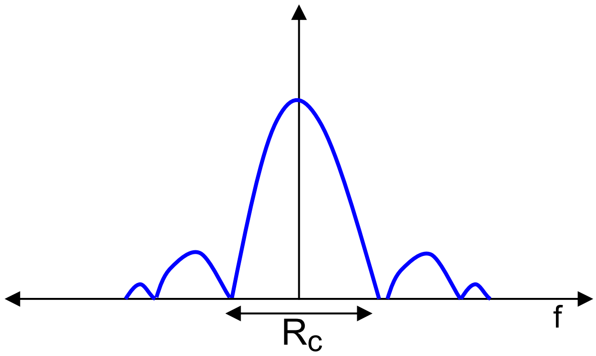


Figure 4.5: BPSK DSSS Spectrum

The bandwidth in DSSS systems is often taken as the null-to-null bandwidth of the main lobe of the power spectral density plot (indicated as 2Rc in Fig. 4.5). The half power bandwidth of this lobe is 1.2 Rc, where Rc is the bit rate. Therefore, the bandwidth of a DSSS system is a direct function of the bit rate. It should be noted that the power contained in the main lobe comprises 90 percent of the total power. This allows a narrower RF bandwidth to accommodate the received signal with the effect of rounding the received pulses in the time domain.

b. Receiver requirements

When designing a receiver, the main goal is to provide the demodulator with a signal with the necessary quality and power level. There are some parameters which determine the limiting values of the quality and power level of the received signal. These parameters are referred to the measurement device (vector signal analyzer) or to the receiver circuit. A good understanding of these parameters is necessary. In the next paragraphs a description of the most important parameters is carried out.

When the signal from the receiver is fed to the input of a vector signal analyzer to be analyzed, the signal has to fulfill some requirements.

The dynamic range of the receiver is its most important performance figure. The dynamic range has different possible interpretations. Some factors have to be taken into account to evaluate the dynamic range.

The dynamic range is defined as the ratio of the largest to the smallest signals in dB which can be handled by the receiver without significant noise or nonlinearity impairments.

The next diagram shows the different parameters to take into account when a good interpretation of the dynamic range is desired:

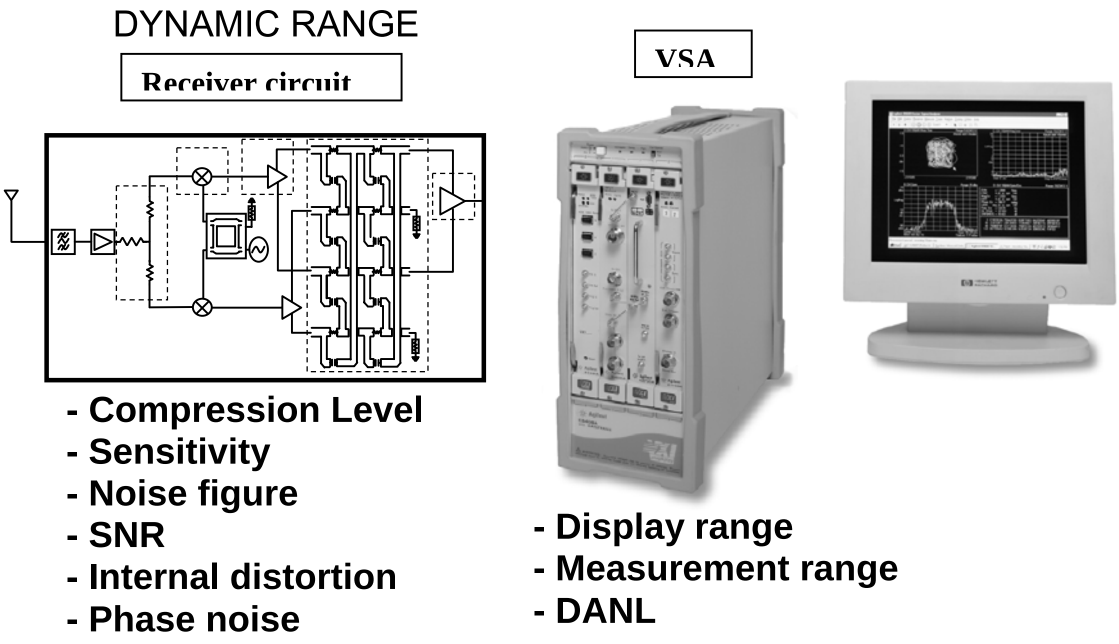


Figure 4.6: Contributions to the dynamic range

As mentioned above, the dynamic range has different possible interpretation. To use the correct interpretation for a determined application, several factors have to be understood; display and measurement range of the measurement device, compression level, sensitivity, noise figure, SNR, internal distortion and phase noise of the receiver circuit.

The influence of each factor on the dynamic range interpretation is described in the following.

The **measurement range** is the difference between the largest and smallest signal that can be measured. Usually, the maximum power level that can be applied to the input of the analyzer without damaging the front-end hardware dictates the largest signal; this is +30 dBm (1 watt) for most analyzers. The **noise floor** of the instrument determines the lower limit of the measurement range. Signals below the sensitivity level of the device are not distinguishable from noise. The noise floor of the instrument is usually called sensitivity or Display Average Noise Level (**DANL**) of the instrument. This parameter is usually given as a function of the resolution bandwidth used. **Resolution bandwidth** (RBW) is the bandwidth of the IF filter which determines the selectivity of the vector signal analyzer. A wide resolution bandwidth is required for wide sweeps while a narrow filter is used for narrow sweeps.

Noise is a broadband signal. Therefore, as the RBW filter is widened, more random noise energy is allowed to hit the detector. This increases the noise level of the analyzer. Therefore, noise specifications must be referenced to the RBW.

The **display range** is the calibrated amplitude range of the display. If small signals need to be measured in the presence of larger signals, the larger signals can often be moved above the upper limit of the display by up to 10 dB by changing the reference level, to place the small signals in the calibrated display range. This has little effect on the accuracy of measuring the smaller signal.

However, the larger signals are limited because the **circuit compression point** must be avoided. The circuit compression level is the maximum power level (the sum of the input power at all frequencies) that can be put into the receiver circuit input without compromising the accuracy of the output signal. When the signal level at the input of the receiver circuit is low enough, the level of the desired IF signal is a linear function of the input and little energy is diverted to distortion. As the input level increases, the transfer function becomes nonlinear because significant energy is lost to the distortion products. At this point the circuit is considered to be in compression. The compression level describes the total input power level below which the circuit compresses the output signal less than 1 dB (1 dB compression point).

Internal distortion in the receiver circuit is one of the factors which determines the dynamic range when measuring distortion products, such as harmonic distortion from a single tone or intermodulation distortion from two or more tones. The internally generated intermodulation and harmonic distortions are a function of the input signal amplitude at the receiver circuit.

If the input level is low enough, there is no need to be concerned about internally generated distortion. In this case, when the signal gets lower, the effects of noise has to be taken into account.

Two types of noise in the receiver contribute to dynamic range: **phase noise** and **sensitivity.**

The sensitivity of the circuit, also called noise floor, is the measure of the signal created from the sum of all the noise sources and unwanted signals within a measurement system. In radio communication and electronics, this may include thermal noise and any other interfering signals. Thermal noise is usually the major contribution to the noise floor. The sensitivity determines the smallest signal that can be measured. The lower limit of the noise floor is theoretically kTB (where k=Boltzmann's constant, T=absolute temperature in degrees Kelvin, and B=bandwidth in Hertz) or –174 dBm, for a noise bandwidth of 1 Hz at room temperature.

For a bandwidth given, the noise floor (P_{NOISE}) is calculated with the next expression:

$$P_{NOISE} = -174 + 10 \cdot \log(BW(Hz)) \tag{4.1}$$

This noise floor determines the minimum power that the circuit is able to realize, but, to know the minimum input power that the receiver needs to demodulate, the knowledge of other parameters is necessary. The noise figure of the receiver NF and the minimum SNR (SNRmin) are necessary to recognize the wanted signal.

The **noise figure** (NF) is a number by which the performance of a radio receiver can be specified. It is a measure of degradation of the signal to noise ratio (SNR), caused by components in the RF signal chain. The **SNR** is the ratio of a signal power to the noise power corrupting the signal.

The maximum allowed receiver noise figure (NF_{max}) at the antenna can be calculated with the next expression depending on the minimum power input level ($P_{in,min}$) and the noise floor (the following expression is deduced in the chapter 2 of the reference "RF microelectronics").

$$NF_{\max} = P_{in,\min} - P_{NOISE} - SNR_{\min} \tag{4.2}$$

While noise floor is the key parameter when measuring two signals that are far apart in frequency, **phase noise** is the key parameter when measuring two signals that are close in frequency (<1 MHz apart). Also called sideband noise, phase noise is caused by the instability of the local oscillator (LO). Any instability in the LO is translated to the output signal through the mixer. The more stable the LO, the lower is the phase noise, assuming a sufficiently stable input signal.

The limiting factors for the dynamic range have been explained above. But until now, the presence of interfering signals has been omitted. The existence of large unwanted signals in the vicinity of the band of interest even after filtering creates difficulties in the design of the following circuits.

A preselection filter can be very useful to reduce even order intermodulation products but odd order intermodulation products can not be reduced by filtering before the final channel selection. For example, when the circuit exhibits nonlinearities, the "intermodulation products" of two strong unwanted signals may appear in the design band, thereby corrupting the reception.

When strong interfering signals are present, they can be the most limiting factor in the dynamic range. When planning to design a receiver, two cases of interfering signals present in the channel have to be taken especially into account because of their consequence on the wanted signal. The first case is due to the presence of a single strong interferer and the second case is due to the presence of two interfering tones close to each other. Because of their importance, a thorough explanation about these effects is done in the next paragraphs.

The reception of a signal can be disturbed by a single strong interferer. The presence of a strong interferer determines several requirements for the receiver, including the total channel selectivity, the minimal required **image-rejection**, the compression point of various blocks, and the phase-noise characteristics of the VCO.

These effects, except the image-rejection, have been explained above. The problem of the image-rejection occurs after down conversion when a single strong interferer appears at a frequency two times the intermediate frequency far apart from the wanted signal.

This problem can be seen in the next graphic:

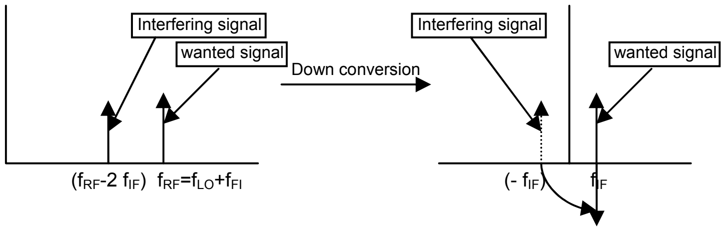


Figure 4.7: The problem of the image rejection

When the received signal is down-converted, both the signal at the frequency of the local oscillator plus the intermediate frequency (wished signal (f_{LO} + f_{IF})) and minus the intermediate frequency (image signal(f_{LO} - f_{IF})) are down-converted to the intermediate frequency because of the mirror properties of the spectrum. This interfering signal can degrade the quality of the signal and complicates its demodulation. Therefore a good value for the image reject ratio is needed. As far as this work is concerned, the problem of the image-rejection is treated with a polyphase filter. A thorough explanation of the image rejection using polyphase filters is given in the paragraph dedicated to the receiver planning.

In the presence of two interference tones close to each other in the frequency domain the most important effect is called third-order intercept point (IP3). In the next paragraph the issue of the IP3 is explained.

The definition of third-order intercept points more frequently used is the intermodulation intercept point, measured by feeding two signals with a small frequency difference into the device-under-test and measuring the output power at the interest frequency.

Third-order intercept point ("IP3") is based on the idea that the device nonlinearity can be modeled using a low order polynomial, derived by means of Taylor series expansion. If the input/output static characteristic of the circuit is approximated as $y(t) = \alpha_1 \cdot x(t) + \alpha_2 \cdot x^2(t) + \alpha_3 \cdot x^3(t) + ...$ and $x(t) = A_1 \cdot \cos(\omega_1 t) + A_2 \cdot \cos(\omega_2 t)$, then the cubic term yields components at 2ω1-ω2 and 2ω2-ω1, either of which may fall in the band of interest. The standard approach to quantifying this effect is to choose A_1=A_2 and, using extrapolation, calculate the IP3 by plotting the output power versus the input power on dB scale. The theoretical calculation of the IP3 can be seen in the next figure:

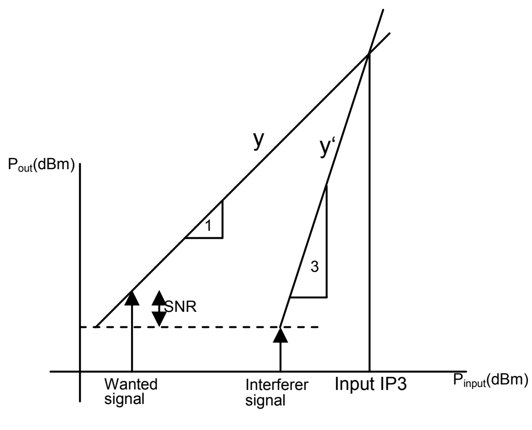


Figure 4.8: Theoretical calculation of the IP3

Two curves are drawn, one for the wanted linearly amplified signal at an input tone frequency, one for a nonlinear product. On a logarithmic scale, "x to the power of n" translates into a straight line with slope of n. Therefore, the linearly amplified signal will exhibit a slope of 1. A 3rd order nonlinear product will increase by 3 dB in power, when the input power is raised by 1 dB. The point where the curves intersect or in other words, the input power that results in equal magnitudes for the fundamental components and the intermodulation products is the intercept point. It can be referred to the input or output power axis, leading to input or output referred intercept point, respectively (IIP3/OIP3).

The two-tone approach has the advantage, that it is not restricted to broadband devices. Both definitions differ by 4.8 dB (10 log10 3).

The IP3 in conjunction with the compression point (P1dB) specifies the linearity of the receiver.

c. Parameters of the vector signal analyzer.

Once the important parameters to take into account are known, the next step consists in determination of the parameters for the planned receiver. The values calculated in this section allow doing the receiver planning and choosing the correct components.

As mentioned above in this chapter, the receiver specification depends on the parameters of the vector signal analyzer. Therefore, firstly the features of the measurement device used to collect the data after the receiver are introduced in this section. Once these parameters are known, the values for these parameters are set in order to carry out the channel measurements. The choice of these values allows going on with the calculation of the minimum requirements of the receiver.

The device used to collect the data from the receiver is the 89600 Vector Signal Analyzer. This device allows different configurations. The configuration used in this work is the 89610 DC-39 MHz Vector signal analyzer. This analyzer provides 39 MHz of analysis bandwidth. 1 or 2 channel configuration can be used. Its 2-channel configuration provides a channel 1 plus j*channel 2 mode for measurements on direct baseband I and Q inputs.

The main component of this configuration is the E1438 analog-to-digital converter which allows a sample rate up to 100Msamples/sec.

Its specification is included in the 89600 Vector Signal Analyzer data sheet. The Agilent 89600 Series vector signal analyzer has two sets of application software: vector signal analysis (VSA) and spectrum analysis. The VSA application software is used to analyze complex signals in the time, frequency and modulation domains. The spectrum analyzer application software emulates a traditional spectrum analyzer, providing fast, high resolution signal magnitude measurements while sweeping across a user-defined frequency span.

The 89600 series VSA has two signal processing modes: baseband and zoom. These two processing modes affect the appearance and the duration of input waveforms displayed by the 89600s. Most 89600 measurements are made with a non-zero start frequency, called the Zoom mode. In these cases, the time domain display shows a complex envelope representation of the input signal – that is, the magnitude and phase of the signal relative to the analyzer's center frequency. This provides powerful capability to examine the baseband components of a signal without the need for demodulation.

Baseband mode refers to the special case where the measurement begins at 0 Hz.

Here, the input signal is directly digitized and the waveform display shows the entire signal (carrier plus modulation), like an oscilloscope would do.

Because very low frequencies will be measured, the signal processing mode chosen is the base band mode. In the following table the most important features for this mode of operation are shown:

Frequency			
			89610A
			(DC – 40 MHz)
Frequency range			
	Vector analysis mode		
		Baseband mode	DC to 40 MHz
Frequency spans	The span defines a frequency range over which the device sweeps across to measure the signal magnitude.		
	Vector signal analyzer application		
		1 channel mode	< 1 Hz to 39.06 MHz
		2 channel mode	< 1 Hz to 39.06 MHz
		Ch1 + j*Ch2 mode	< 2 Hz to 78 MHz
Frequency points	The number of frequency points refers to the number of samples of the signal which are stored during a time interval corresponding to the resolution bandwidth used. It has to be less than or equal to the number of samples which the device is able to collect during this time.		
	Vector signal analyzer application		
		Calibrated points	51 – 102,401

		Displayable points	51 – 131,072
Resolution Bandwidth Filtering			
			89610A
			(DC – 40 MHz)
RBW range	The resolution bandwidth (RBW) is the minimum bandwidth in which the measurements have enough resolution. Therefore it must be equal or higher than the bandwidth of the signal transmitted. The RBW defined the selectivity of the measurements as well. Therefore but must be chosen as slow as possible. The range of available RBW choices is a function of the selected frequency span and the number of frequency points.		
	Vector signal analyzer application		< 1 Hz to 10 MHz
Amplitude			
Input			
	Full-scale range		
		Baseband mode	–31 dBm to +20 dBm in 3 dB steps
	Maximum safe input level		+24 dBm, ±5 VDC
	Input channels		
		Standard	1
		Optional	2 baseband
	Nominal impedance		50 ohms
Dynamic range			
	Input noise density	Range ≥ –30 dBm.	
		Baseband mode (> 0.1	< –121 dBfs/Hz

		MHz)	
	DANL	Most sensitive range.	
		Baseband mode	< –151 dBm/Hz

Table 4.1: Vector signal analyzer parameters

In the 89600 VSA applications, measurements are based on time records. A time record is a block of samples of the signal waveform from which time, frequency and modulation domain data is derived. Time records have the following characteristics:

$$\text{Time record length (main time)} = \frac{\text{(Number of frequency points - 1)}}{\text{span}} \tag{4.3}$$

$$\text{Time sample resolution} = \frac{1}{\text{k} \cdot \text{span}} \tag{4.4}$$

where:

k = 2.56 for time data mode set to baseband

k = 1.28 for all other modes (default) including zoom

Span = currently selected frequency span

In time capture mode the 89600 VSA application captures the incoming waveform gap-free into high-speed time capture memory. This data may then be replayed through the analyzer at full or reduced speed, saved to mass storage, or transferred to another application software.

$$\text{Time sample resolution} = \frac{1}{\text{k} \cdot \text{cardinal span}} \tag{4.5}$$

where:

k = 2.56 for time data mode set to baseband (89610A only)

k = 1.28 for all other modes (default) including zoom

$$\text{Cardinal span} = \text{max. span/2n, for n} = 0 \text{ to } 17 \tag{4.6}$$

During time capture, the analyzer is internally set to the cardinal span that equals or exceeds the currently displayed frequency span. The cardinal span is defined by Maximum Span/2n where n is an integer.

In this point, the most important parameters of the VSA are shown. Another very important parameter is the sample rate of the VSA. The data sheet of the VSA doesn't give this parameter, but this parameter can be calculated by using its parameters. An explanation of the way to calculate this parameter is given as follows.

When the device isn't in recording (or time capture) mode, the sample rate is determined by the span selected and two different cases have to be considered. When zoom time is used (i.e. the start frequency is not zero Hz), the equation is:

$$\text{Sample Rate (Hz)} = 1{,}28 \cdot \text{User Span.} \tag{4.7}$$

When baseband time is used (i.e. the start frequency is zero Hz), the relationship is:

$$\text{Sample Rate (Hz)} = 2{,}56 \cdot \text{User Span.} \tag{4.8}$$

However, in recording or time capture mode, the sample rate is always determined by these equations:

$$\text{Sample Rate (Hz)} = 1{,}28 \cdot \text{Cardinal Span.} \tag{4.9}$$

When baseband time is used, the relationship is:

$$\text{Sample Rate (Hz)} = 2{,}56 \cdot \text{Cardinal Span.} \tag{4.10}$$

A cardinal span is defined by:

$$\text{Cardinal Span} = \frac{\text{Maximum Span}}{2\text{n}} \tag{4.11}$$

Where n is an integer.

d. Receiver specifications

Until this point some factors which determine the receiver specifications have been explained. Concretely, the receiver specifications are determined by the choices of the parameters of the vector signal analyzer, the limits imposes by the wireless mobile radio propagation channel and the minimum requirements of the receiver outlined in the previous chapter.

In the next paragraphs values that fix the limits for the receiver specification are explained and calculated. A diagram with these values is placed at the end of this section to give an overview of the required values. Some parameters of the vector signal analyzer must be fixed before the definition of the receiver specifications. These parameters are fixed along this section when they are needed. A short explanation of the reasons of these choices is given when these values are chosen. At the end of this paragraph the receiver specifications which fit best the requirements of the receiver to design are defined. These specifications pave the way to the receiver planning.

As mentioned in the chapter 2.1, the wireless communication channel sets several limits in a transmission. In the next paragraphs these limits are calculated.

As explained in the mentioned chapter, the **small scale effects** of the wireless mobile channel set different limits in the symbol time. The delay time refers to the time-spreading manifestation which results from the imperfect channel impulse response and imposes a lower limit on the symbol time. The transmission time is related to the time-varying nature of the channel caused by relative motion between a transmitter and receiver, or by movement of objects within the channel and imposes an upper limit in the symbol time. Concretely, the transmission time defines the maximum time the signal can be transmitted and stored while the channel remains nearly constant.

For an indoor scenario, based on the published information on radio propagation and recently studies about this topic, a sufficiently slow value for the transmission time (TT) is 10 ms.

The next lines are dedicated to evaluate the parameters which the receiver needs to work with transmitted signals which fulfill the mentioned condition.

Firstly some characteristics of the transmitted signal have to be known. The first characteristic that has to be known is the modulation used.

As explained in the section 4.a. the signal received by each receive antenna is the sum of different signals. Therefore the transmitted signal must use a modulation scheme enabling the receiver to separate both signals. As explained in this chapter, a spread spectrum technique can be used to achieve it. The modulation used is a BPSK.

BPSK conveys data by changing, or modulating, the phase of a reference signal. Concretely, BPSK uses two phases which are separated by 180° and modulated at 1bit/symbol. It is very robust and takes serious distortion to make the demodulator reach an incorrect decision. For more information about the modulation scheme, the reader is referred to the reference “Signal processing for estimation of narrowband MIMO transmission channel”.

The mentioned reference process the data stored in the VSA to estimate the capacity of the channel. Due to the algorithm used in this work, a signal of 1000 symbols is transmitted to make the recognition of the symbols transmitted easier.

The bandwidth of the symbol transmitted is chosen to avoid that the duration of the signal transmitted exceed the time in which the channel remains nearly constant.

In order to measure the transmitted signal when the channel remains nearly constant, the time record length, when measuring the block of thousand symbol, don’t have to exceed 10 milliseconds.

In the next paragraph, a bandwidth of 300 kHz (150 kHz base band bandwidth) for the transmitted signal is checked to evaluate if this signal exceeds the maximum time record length.

The first step consists in calculating the duration of the transmitted signal.

$$Symbol\_time = \frac{1}{150 \cdot 10^3} = 6{,}66 \cdot 10^{-6} seg \tag{4.12}$$

$$Duration\_transmitted\_signal = Symbol\_time \cdot 1000 = 6{,}66 mseg \tag{4.13}$$

To calculate the required time record length, to set a frequency span is necessary. The span depends on the number of samples per symbol to be collected. To keep balance between reducing the possibilities of picking one big interferer frequency and obtaining a high sample rate, a quite high number of samples per symbol are achieved with a span of 2 MHz.

The sample rate of the VSA and the number of samples per symbol can be calculated from the expression …

$$span = 2MHz \Rightarrow Sample\_rate = span \cdot 2{,}56 = 5{,}12 MSamples/seg. \tag{4.14}$$

To know the sample rate allows calculating the number of samples per symbol:

$$Samples = Symbol\_time \cdot Sample\_rate = 34{,}13 Samples \tag{4.15}$$

As mentioned above, the number of frequency points fulfils the next relation:

$$Number\_of\_frequency\_po\text{int}s \geq Samples\_per\_symbol \tag{4.16}$$

With these values the time record length can be calculated with the following expression:

$$\text{Time record length} = \frac{(\text{Number of frequency points - 1})}{\text{span}} = \frac{(34{,}13 \cdot 1000 - 1)}{5 \cdot 10^6} \tag{4.17}$$
$$= 6{,}82 mseg$$

As can be seen, this time is less than the transmission time and therefore an invariant behavior of the channel during a single transmission of 1000 symbols can be expected.

The symbol bandwidth determines the minimum resolution bandwidth which can be selected to obtain a good resolution of the received signal.

Typical values used in engineering for the delay time are in indoor channels approximately 50 ns whereas in outdoor microcells approximately 30µsec.

Due to the long duration of each transmitted symbol (6.6 µsec.), the lower limit imposed by the delay time in indoor channels (50 ns) is far below the duration of a transmitted symbol. In other words, all the received multipath components of a symbol are expected to arrive within the symbol time duration. Therefore, as explained in chapter 2.1, there is no channel-induced ISI distortion and the only effect in this case is loss in SNR.

As explained in a previous chapter, another effect of the wireless mobile channel set some restrictions in the receiver requirements. The so-called "**large scale effect**" refers to the losses in the channel due to the distance between transmitter and receiver.

Receiver parameters like gain and losses in the receiver circuit are limited by the power received by the antenna. Knowing the losses in the wireless channel allows fixing this receiver parameters. Moreover, other factors have to be taken into account. Some of these factors are the maximum power transmitted (regulated by the authorities), the minimum signal to noise ratio (SNR_{min}) required, the sensitivity of the receiver circuit (Noise floor) and the VSA (DANL) and the maximum noise figure (NF_{max}) of the receiver circuit. Knowing this enables to predict the required performance figure of each block and the required gain in the front-end.

These factors were explained during the section "receiver requirements". The values of these factors are calculated in the following. After that, the expected losses in the wireless channel are calculated.

To calculate the NF_{max}, the DANL has to be known. By using the values given in the previous chapter, the sensitivity of the VSA (DANL) is given as follows as a function of the resolution bandwidth as $DANL = -151\,\text{dBm/Hz}$.

As mentioned before, the signal bandwidth used in this work is 300 kHz. This implies that the minimum RBW allowed is 300 kHz. By using this value, the DANL is given as:

$$DANL = -151\,\text{dBm} + 10 \cdot \log(300 \cdot 10^3) \approx -96{,}23dBm \qquad (4.18)$$

If such signal is present at the input of the VSA, it can be recognized by the VSA. But to demodulate the signal collected by the VSA, a higher level of this signal is necessary. This level is determined by the parameter "minimum signal to noise ratio" (SNR_{min}). This parameter defines the minimum relation between the signal level and the noise level to allow the demodulation algorithm to demodulate the signal and recognize the transmitted symbol.

The SNR_{min} depends on the maximum of the bit error rate (BER) to achieve and the modulation scheme used. As the maximal BER is unknown, the SNR_{min} is unknown as well. However, the published information on similar works has been reviewed to obtain a safe value for this parameter. After that, 20 dB SNR_{min} is considered a safe value for this parameter compared with the required values in the published information reviewed.

For the purposes of this work, the receiver must be able to demodulate -90 dBm signal (P_{min_input}).

By using the expression given in the previous section, the maximum allowed receiver noise figure for a 300 kHz resolution bandwidth is calculated as follows:

$$NF = P_{in,\min} - noise\_floor - SNR_{\min} \tag{4.19}$$

$$NF = -90 - (-174 + 10\log(300 \cdot 10^3)) - 20 = 10{,}77dB \tag{4.20}$$

The knowledge of the minimum power at the input of the receiver circuit P_{input_min}, the minimum signal to noise ratio SNR_{min} and the sensitivity of the VSA S_{VSA} allow knowing the necessary gain $Gain_{min}$ in the receiver front-end. This can be calculated with the following expression:

$$Gain_{\min} = S_{VNA} + SNR_{\min} - P_{input\_\min} = -96{,}23 + 20 + 90 = 13{,}77dB \tag{4.21}$$

To carry out the estimating path loss in the wireless channel two kinds of information sources can be used, empirical models and experimental results.

The first way to find some values of the path loss is to calculate it with empirical indoor propagation models.

As explained in the chapter dedicated to the fading in wireless mobile channels, the next expression can be used to calculate the path-loss.

$$\overline{L}_P(dB) = L_s(d_0)(dB) + 10 \cdot n \cdot \log(d / d_0) + X_s \tag{4.22}$$

Where, typical values for Xs are as high as 6-10 dB or greater.

In presenting the results of experimental data in a multipath environment, it is essential to include a good description of the areas under test. However, for the wireless channel, some compromises must be made. This is because the received signal statistics depend on many factors which vary significantly for different types of buildings. A detailed description of a building under test would make the classification of similar types of buildings difficult. Hence, it is usually attempted to provide a classification which, to some extent, disregards the detailed structure of any particular building. Such a classification results in a wide variation of the statistics of the received signal over buildings belonging to a same category, but can be a good reference for estimating an approximate value of the path loss in our conditions. For the purposes of this work the scenario can be classified as an “Office buildings in suburban areas”.

Examination of the published experimental results of indoor radio wave propagation studies shows values of the parameter n (gradient) of smaller than 2 for corridors and large open indoor areas and values as large as 6 for buildings with metal walls. In residential areas, offices and manufacturing floors, gradient of 2 to 3 are usually recommended for computer simulation. Larger values of n can be expected when obstructing objects are present.

In the following table, some results for different values of the distance are presented for a gradient value of 3 because this is the value recommended for offices in several documents and also is the decay exponent calculated in some measurements found in references like “Comparative Indoor RF Channel Soundings at 2, 5 and 17 GHz”.

By using the mentioned expression, the losses expected are:

$$\overline{L}_P = \frac{d}{d_0}^n \xrightarrow{DECIBELS} \overline{L}_P(dB) = L_s(d_0)(dB) + 10 \cdot n \cdot \log(d/d_0) \pm X_s \qquad (4.23)$$

d	$L_s(d_0)$	$10 \cdot n \cdot \log(d/d_0)$	$\overline{L}_P$
(m)	(dB)	(dB)	(dB)
1	40,223	0	40,223 $\pm X_s$
10	40,223	30	70,223 $\pm X_s$
20	40,223	39,031	79,254 $\pm X_s$
30	40,223	-44,313	84,536 $\pm X_s$
40	40,223	-48,062	88,285 $\pm X_s$
50	40,223	50,969	91,192 $\pm X_s$
100	40,223	59,777	100,223 $\pm X_s$

Table 4.2: Path loss

As was explained in a previous chapter, a typical value for Xs is 6-10 dB or greater when some obstacles are present. A usual margin in this last case, where some events occur, for example motion of pedestrian, is 40 dB.

At this point the time and power specifications of the receiver are fixed. As mentioned when the receiver requirements were explained, strong interference signals can be present doing the receiver requirements stricter. As explained before, the most important parameters regarding these interfering signals are the IIR (“Image reject ratio”) and the compression when a single strong interference is present, and IP3 when two tones occur close to each other in the nearby the wanted signal.

As explained in the chapter dedicated to the architectures, the IRR depends on the δG and δφ (gain and phase mismatch) between the I and Q paths (including any quadrature error of the LO signals). This relation can be seen in the figure below.

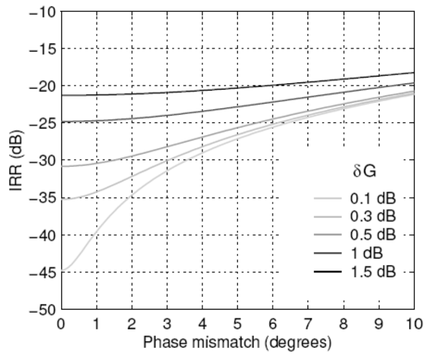


Figure 4.9: IRR as a function of the phase and gain mismatch

Therefore, this value depends on the symmetry and good match of all components in the receiver. To reach a high value of this parameter is one of the most important issues of this work.

Due to the characteristics of components used a phase mismatch of 3 degrees and 1 dB can be achieved. Therefore an IRR of approximately -25 dB can be expected.

Due to the fact that the band used in the transmission is a free band, two interferer signals close to each other can be expected. The intermodulation products of these interferer signals can disturb the wanted signal. Values up to -40dBm can be expected for these interferers.

As mentioned above, a -90 dBm (P_{min_input}) in the input of the receiver circuit is the minimum power signal level that must be demodulated. As SNR must be 20 dB in order to demodulate the signal, the next diagram shows the procedure to obtain the value of IP3 in the presence of two interference signals of -40 dBm:

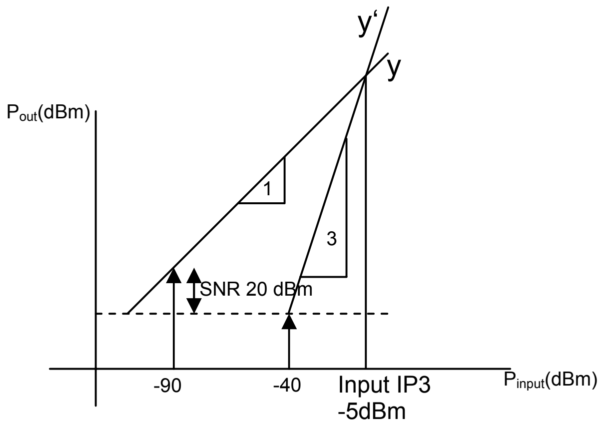


Figure 4.10 Graphical calculation of the IP3

Fig. 4.10 shows the power output of the receiver circuit as a function of the power input. The procedure followed to calculate the IP3 is explained in the chapter “receiver requirements” and the IP3 observed is -5 dBm.

Mathematically, from the values observed in the diagram, the IP3 can be calculated as follows:

$$\left.\begin{aligned} y &= a + 10\cdot x \\ y' &= a' + 30\cdot x' \end{aligned}\right\} x_1 = -110; x_2 = -40 \Rightarrow y = y'\,(y = y' = x + Gain - SNR) \Rightarrow$$

$$\left.\begin{aligned} y &= 1100 - 110 + Gain - SNR + 10\cdot x \\ y' &= 1200 - 110 + Gain - SNR + 30\cdot x' \end{aligned}\right\} \xrightarrow{y=y';x=x'} x = -5dBm$$

Another simpler way to calculate the IP3 but more difficult to observe graphically is by the next expression obtained from the reference xxxx:

$$\begin{aligned} &-110dBm + x\cdot 30dB = -40dBm + x\cdot 10dB \Rightarrow x = 3{,}5 \\ &\Rightarrow IP3,in = -40 + 35 = -5dBm \end{aligned} \tag{4.23}$$

A summary of the values fixed during this section can be seen in the next diagram:

Time specifications

Max time record length 10 ms.

Signal transmitted
- Symbol length = 1000 symbols
- BW = 300 kHz
- Signal time length = 6,66 msec
- Symbol time length = 6,66 µsec

VSA parameters
- Span = 2 MHz } Sample rate = 5,12 MSa/sec
- RBW = 300 kHz

Samples per symbol = 34,09 samples
Time record length = 6,82 msec

Power specification

VSA parameters
- Span = 2 MHz } Sample rate = 5,12 MSa/sec
- RBW = 300 kHz
- DANL = -151 dBm/Hz

RBW, DANL } DANL = -96,23 dBm

SNR_{min} = 20 dBm
$P_{min\ input}$ = -90 dBm

} $Gain_{min}$ = 13,77

} RF receiver NF = 10,77 dB

Wireless channel parameters
- Noisefloor = -174 dBm/Hz
- Channel losses
  - d = 1m → $\overline{L}_p(dB) = 40{,}223 \pm X_s$
  - d = 10m → $\overline{L}_p(dB) = 70{,}223 \pm X_s$
  - d = 50m → $\overline{L}_p(dB) = 91{,}192 \pm X_s$

} Pin = -50
Pin = -80
Pin = -101

$P_{transmitted}$ = 0 dBm

Interference = -40 dBm → $IP3_{max}$ = 10 dBm
IRR_{min} = 25 dB

The red value indicates that in these conditions (Transmitted power = 0 dBm; d = 50 m), the receiver is not able to demodulate the signal.

e. Receiver planning

Until now, all the efforts have been directed to provide the RF designer with the necessary parameters to design the receiver. In this section this parameters are taken into account and a planning of the design is done. At the end of this section, the designed receiver is completely defined.

The purpose of this section is to concrete all components used and explain the choice of these components as well as their features.

When planning a receiver design, the receiver layouts is an important stage. This stage depends on several factors as the transmission line, the substrate, the operation frequency, etc. Due to the importance of the substrate, its main features as well as its influence on some parameters of the transmission line are shown in this section. Microstrip transmission lines are used to carry the signal. When the substrate features are shown, the different parameters of this line are calculated.

Firstly a short explanation of the complete receiver circuit is done in order to provide the reader with a general overview of the designed receiver. After that, each block of the receiver is explained in detail.

When designing a receiver, the goal is not only to fulfill the receiver requirements, but also to do it on the best way for the future application. There are a lot of possibilities to choose the components of a receiver fulfilling these requirements. Therefore, in this work, simulations of the behavior of different components have been done to compare them and to evaluate which components fits best in the receiver. Simulation of critical components of the circuit have been done to check their theoretical features and to probe if these components match proper the desired receiver features. The simulations have been carried out by using the program ADS (“Advanced Design System”).

Advanced Design System (ADS) is an electronic design automation software system produced by Agilent EEsof EDA. This program has been chosen because it supports every step of the design process, as schematic capture, layout, frequency-domain and time-domain circuit simulation, and electromagnetic field simulation, allowing the engineer to fully characterize and optimize an RF design without changing tools.

Some of the schematics and layouts implemented as well as the simulations of the critical components are included in this chapter.

i. Complete receiver

In Fig 4.11 a diagram of the complete receiver can be seen.

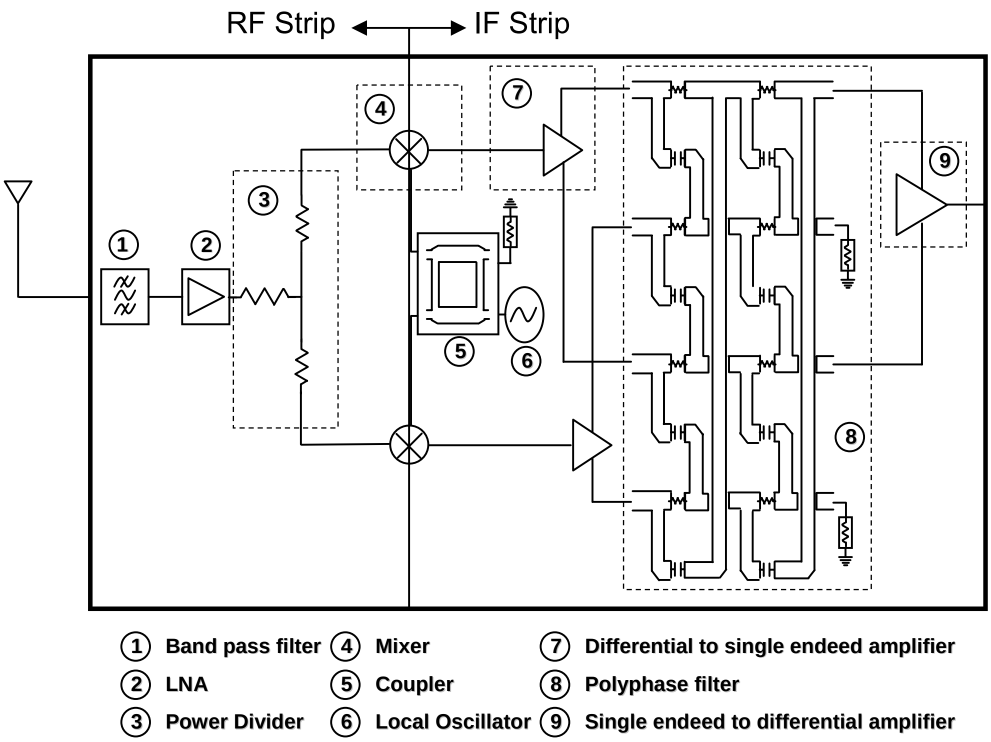


Figure 4.11: Complete receiver

A short summary about the operation of the circuit was given in the chapter "Proposed Design". In this section, the most important information given in that chapter is reminded and more concrete information is added since at this point more information about the receiver is known.

As can be seen in the figure, the receiver architecture is a low IF architecture. The lowest IF as possible has been chosen because of the several advantages of a low intermediate frequency, as simplification of the IF strip and other advantages seen when advantages and disadvantages of a low IF were compared.

In the chapter dedicated to the receiver specification, a bandwidth of 300 kHz for the signal was checked to probe that it fulfils the time channel requirements. An intermediate frequency of two times the baseband bandwidth of the signal (2 times 150 kHz) is the minimum recommendable value for the IF. For this reason an intermediate frequency of 300 kHz is chosen.

As previously mentioned, the symmetry of the complete receiver is the most important parameter taken into account when planning the receiver to avoid degradation of the signal. The symmetry of the circuit is very important due to the fact that the image rejection of the signal is done by combining the two paths in which the signal received is devided by means of a polyphase filter, and any mismatch in the circuit provokes reduction in the image reject ratio. Therefore, all the components of the circuit are chosen with symmetric properties.

Some filters can be placed onto the circuit to suppress frequency component out of the desired band. For example a preselection band pass filter can be placed in the first receiver stage to reject out-of-band components. This filter can be designed with relaxed requirements to reject possible strong interfering signals that can saturate the next receiver stages. Another filter can be placed in the IF strip to reduce possible harmonics generated by nonlinear components. As mentioned above, the intermediate frequency is very low. Therefore the design of such a filter has relaxed requirements.

In the next sections, each block is explained in detail.

Firstly a short introduction of the characteristics of the transmission line is given.

ii. Transmission line

As mentioned above, microstrip transmission lines are used to carry the signal. To calculate its features, the properties of the substrate must be known. The substrate used in this work is a fiberglass-based FR4 board with the following parameters:

Dielectric constant (Er) 4.3

Thickness (H) 1.57 mm.

Metal thickness (T) 0.35 mm

Dielectric loss tangent (TanD) 0.022.

Relative permeability (Mur) 1.0

Heigh of upper shield (Hu) 3.9e34 mil

Metal conductivity (Cond) 4.1e7

Metal roughness (Rough) 0

When knowing the substrate features, the parameters of the microstrip line can be calculated for a determined frequency. For this purpose ADS contains a tool called “Linecalc” to analyze a wide variety of different transmission line types. With “Linecalc” the width of a transmission line can be calculated to obtain a determined characteristic impedance of the line.

To introduce the central frequency of operation is necessary too, but this parameter doesn’t have influence in the width of the line. The receiver circuit has two different frequencies. In the first stage, the circuit works with a 2.45 GHz signal and after the mixer, at 300 KHz. At high frequencies, the impedance of the line is a critical factor. Therefore the width of the line is calculated to achieve 50 Ω characteristic impedance at 2.45 GHz centre frequency. This parameter is calculated as follows:

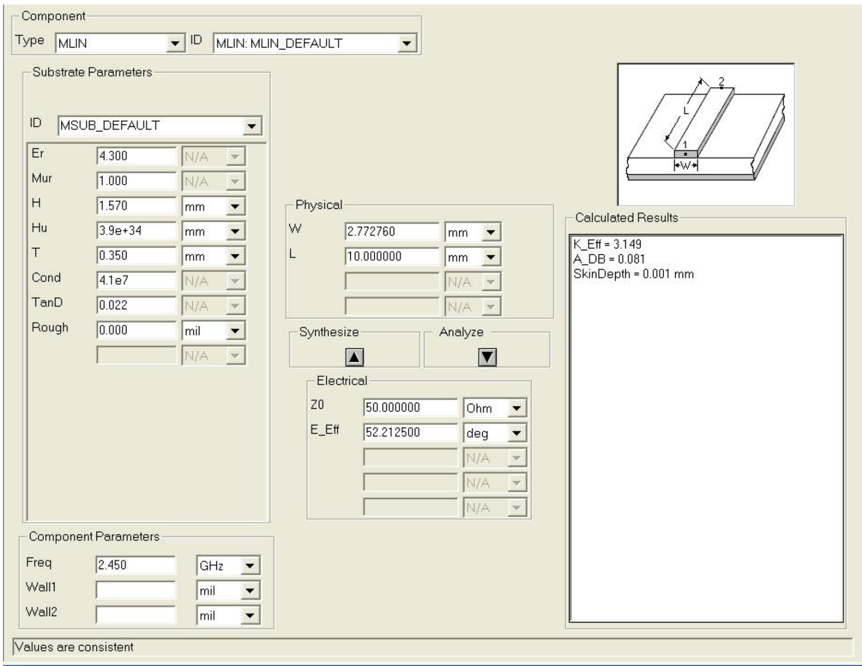


Figure 4.12: Calculation of the TL width by using the tool Linecalc

As can be seen, the width of the microstrip line is 2,772 mm. The parameters in the picture doesn't mentioned until now (for example electrical length (E_EFF)) doesn't have influence in the line width.

Regarding the physical length of the line, a value as small as possible and always less than quarter-wavelength have to be used to avoid impedance transformation. Therefore an electrical length less than 90° (=λ/4) has been used.

For the second stage of the circuit (f = 300 kHz) the frequency is quite low. Therefore the width of the microstrip line is not so critical and the width of the line used has been reduced.

Transmission lines are used as well to feed the components with DC voltage. In this case the width of the line depends on the current. The maximum current flowing in the circuit is needed to feed the amplifiers, therefore the minimum width of the line is calculated for this line. For this purpose a calculator has been used. This calculator used can be found in the reference of this chapter. The parameters needed are the current, the material thickness and the temperature rise (how much hotter the trace will get with current flowing in it compared to without. 10 degrees is a very safe number to use for just about any application).

The results obtained can be seen in the next capture:

Inputs:

Current	0.065	Amps
Thickness	1.57	mm

Optional Inputs:

Temperature Rise	10	Deg C
Ambient Temperature	25	Deg C
Trace Length	20	mm

Results for Internal Layers:

Required Trace Width	0.000402	mm
Resistance	0.560	Ohms
Voltage Drop	0.0364	Volts
Power Loss	0.00237	Watts

Results for External Layers in Air:

Required Trace Width	0.000154	mm
Resistance	1.46	Ohms
Voltage Drop	0.0948	Volts
Power Loss	0.00616	Watts

As can be seen, the minimum required trace width is quite low. The width used in this work for these lines is wider, therefore no problems are expected.

iii. Low noise amplifier

As can be seen in figure 4.11 the low noise amplifier is used after the preselection filter in the first stage of the receiver circuit. As explained in the previous chapter, the power level of the signal is reduced because of the losses in the wireless channel. The attenuated signal reaches the antenna with very low power and is also attenuated by distinct components in the receiver circuit. Since the receiver system require a certain power level to be able to process the signal, the first stages of the circuit must amplify the signal received. Therefore the function of this amplifier is to provide the circuit with enough signal power.

Other factors must be taken into account as well. For example the noise figure of this component is very important. The first stage of the circuit is the most important when calculating the noise figure. Therefore a low noise figure amplifier has to be used. Other important factors of the amplifier are shown in this section.

Also amplifiers have a limited range of operation frequencies. The amplifier must be chosen with a range of operation frequency which includes the bandwidth of the signal at the operation frequency of the circuit.

Therefore the most important requirements of the amplifier are its gain and its noise figure. In this section, taking into account these parameters, the amplifier is chosen to fulfil the receiver requirements.

The model chosen, its more important characteristics and the application circuit used are shown as well.

The amplifier chosen is the monolithic surface mount amplifier model ERA-5+. Concretely two of these amplifiers are placed in the first stage of the circuit.

Figure 4.13: Amplifier model ERA 5+

Its main features are summarized in the next table:

Model Number	Frequency Range		Gain	Max. PowerOutput @1 dB comp.	N.F.	IP3	VSWR (:1)		Device DC Operating Power	
	Low (MHz) High		(dB)	(dBm)	(dB)	(dBm)	In	Out	Voltage(V)	Current(mA)
ERA-5+	DC	4000	18.5	18.4	4.3	32.5	1.3	1.2	4.9	65

Table 4.3: Features of the model ERA 5+

As can be seen in table 4.3, this amplifier fulfils all the required features. Some of these features are compared with the minimum requirements in the next lines.

Its frequency range of operation covers from DC to 4 GHz. This amplifier must treat with a signal of 300 kHz bandwidth and 2.45 GHz centre frequency. Therefore the frequency range of this amplifier is appropriate.

The gain of the amplifier is typically 18.5 dB. The required gain of the receiver front-end is 13.77 dB. Due to the losses of the circuit, one amplifier is possibly not enough to cover the requirements. Two of these amplifiers are placed in the circuit to be well above the requirements.

The noise figure of the amplifier is 4.3 dB. The maximum noise figure of the receiver front-end is 10.77 dB. As the amplifier is the component which has the most influence on this parameter, the margin given is enough.

The IP3 of the amplifier (32.5 dBm) is quite higher than the minimum IP3 calculated for the receiver front-end (-5 dBm).

In the next lines some details of the amplifier given in its data sheet are shown. The reader can find more details in the data sheet of this component.

Firstly the S-parameters of the amplifier versus the current (considering the RF input the pin 1 and the RF output the pin 2) are given:

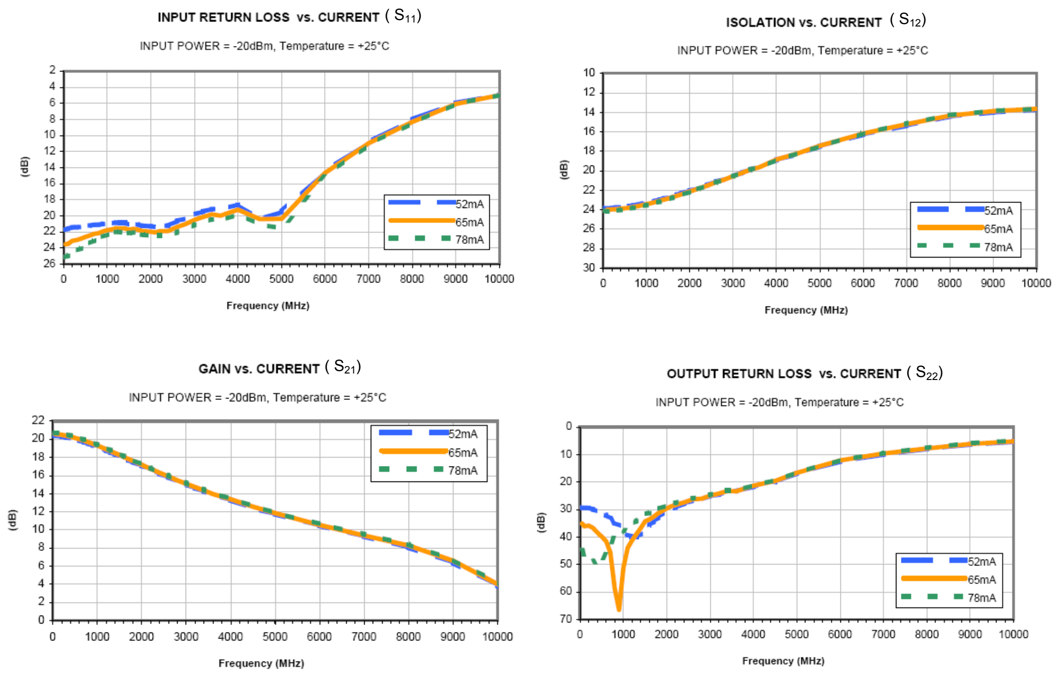


Figure 4.14: S-parameters of the amplifier model ERA 5+

The simplified schematic and pin description of the amplifier model ERA-5+ can be seen as follows:

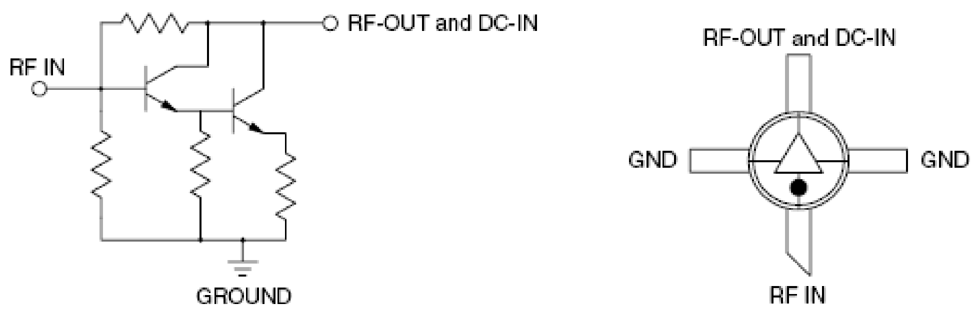


Figure 4.15: Simplified schematic and pin description of the amplifier model ERA 5+

The pin number 1 is the RF input and requires the use of an external DC blocking capacitor chosen for the frequency of operation. The pin number 3 is the RF output and bias pin. DC voltage is present on this pin. Therefore a DC blocking capacitor is necessary for proper operation. An RF choke is needed to feed DC bias without loss of RF signal due to the bias connection. The rest of pins are connections to ground.

Finally the schematic and a picture of the application circuit are shown:

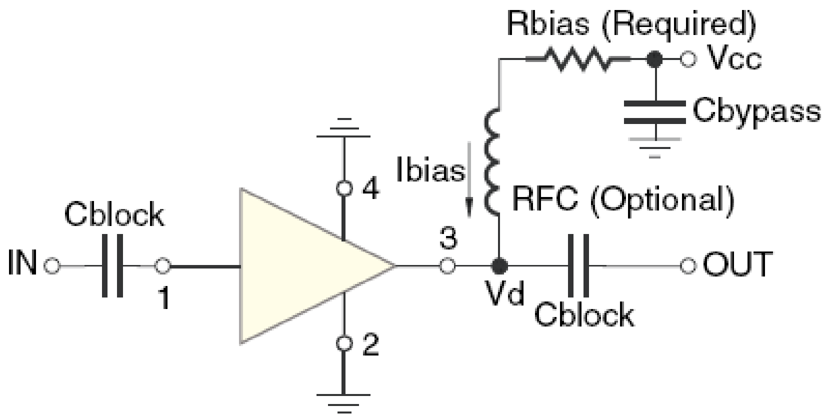


Figure 4.16: Application circuit of the amplifier model ERA 5+

Some considerations have to be done about this circuit. The first one is the resistor values (Rbias) for optimal biasing. The device operating voltage must be at least 4.9 as can be seen in its specifications. The value of the Vcc used to feed the circuit is set to 6.2 V. Therefore the value of the Rbias can be calculated as follows:

$$R_{bias} = \frac{V_{cc} - V_d}{I_{bias}} = \frac{6{,}2 - 4{,}9}{0{,}065} = 20\Omega \tag{4.24}$$

Another consideration is the RF Choke used. As can be seen in the application circuit, a resistor and a RF Choke are used in series with the DC supply. The purpose of the RF choke is to minimize the RF loss caused by the resistor. Performance characteristics such as gain, return loss, IP3, and power output are improved as well when using this RF Choke.

iv. Power divider

Once the signal is amplified, it must be divided in two paths (I and Q paths). To divide the power into two paths, different ways can be followed. The devices with these functions are called power dividers. They have three or more ports and usually divide the power in equal parts. In a three port net, the S-parameters matrix is defined as follows:

$$[S] = \begin{pmatrix} S_{11} S_{12} S_{13} \\ S_{21} S_{22} S_{23} \\ S_{31} S_{32} S_{33} \end{pmatrix} \tag{4.25}$$

Some properties must been taken into account when a power divider is chosen. Three properties are the most important, the losses, the port match and the reciprocity of its S-parameters matrix. Since when the matrix is a 3x3 matrix (the net is a 3 port net) a net with reciprocal matrix and without losses can't have all its port matched, a trade-off between these three parameters has to be looked for.

In this work a resistive power divider is used. With this net, a reciprocal matrix with all its ports matched is achieved, but some losses are introduced. A diagram of this circuit can be seen below:

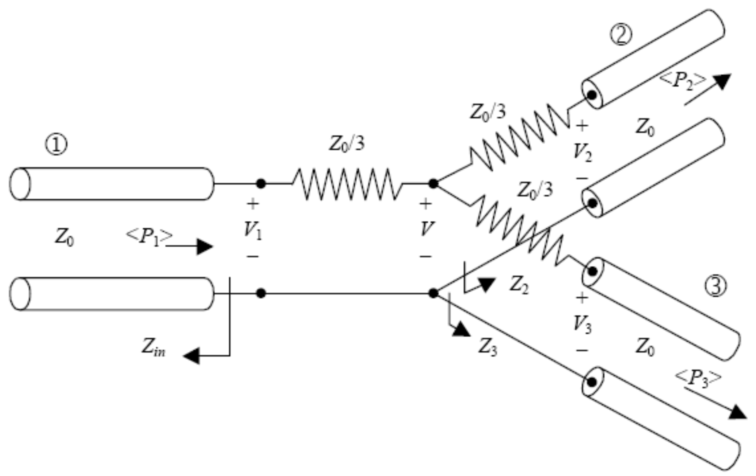


Figure 4.17: Scheme of a resistive power divider

As can be seen, this net is a T-union net where resistive components are added in its branches to get a match in all its ports. A mathematical analysis of this net is presented in the following.

If the ports 2 and 3 are finished with load Z0, the branches 2 and 3 present the following impedances:

$$Z_2 = Z_3 = \frac{Z_0}{3} + Z_0 = \frac{4 \cdot Z_0}{3} \tag{4.26}$$

The impedance at port 1 is:

$$Z_{in} = \frac{Z_0}{3} + (Z_2 \| Z_3) = Z_0. \tag{4.27}$$

Due to the fact that the net is symmetrical:

$$S_{11} = S_{22} = S_{33} = 0. \tag{4.28}$$

In addition:

$$V = V_1 \cdot 2 \cdot \frac{Z_0/3}{Z_0/3 + 2 \cdot Z_0/3} = \frac{2}{3} \cdot V_1 \Rightarrow V_2 = V_3 = \frac{Z_0}{Z_0 + Z_0/3} = \frac{3}{4} \cdot V = \frac{V_1}{2}. \tag{4.28}$$

Therefore:

$$S_{21} = S_{31} = S_{23} = \frac{1}{2}. \tag{4.29}$$

Then, the resulting S-parameters matrix is:

$$[S] = \begin{pmatrix} 0 & 1/2 & 1/2 \\ 1/2 & 0 & 1/2 \\ 1/2 & 1/2 & 0 \end{pmatrix} \tag{4.30}$$

Therefore, the losses can be calculated as follows:

The power at the input is:

$$P_{in} = \frac{|V_1|^2}{2 \cdot Z_0} \tag{4.31}$$

and at the output is:

$$P_2 = P_3 = \frac{1}{2 \cdot Z_0} \cdot \left|\frac{V_1}{2}\right|^2 = \frac{1}{4} \cdot P_{in} \tag{4.32}$$

Therefore the losses are

$$P_{dis} = P_{in} - P_2 - P_3 = P_{in} - \frac{1}{2} P_{in} = \frac{1}{2} P_{in} \tag{4.33}$$

That is, half the power is dissipated (Pdis) in the resistive elements.

Due to its simple mode of operation, this component has not been checked stand-alone. The program ADS has been used to simulate it. The schematic implemented and the results obtained are presented below:

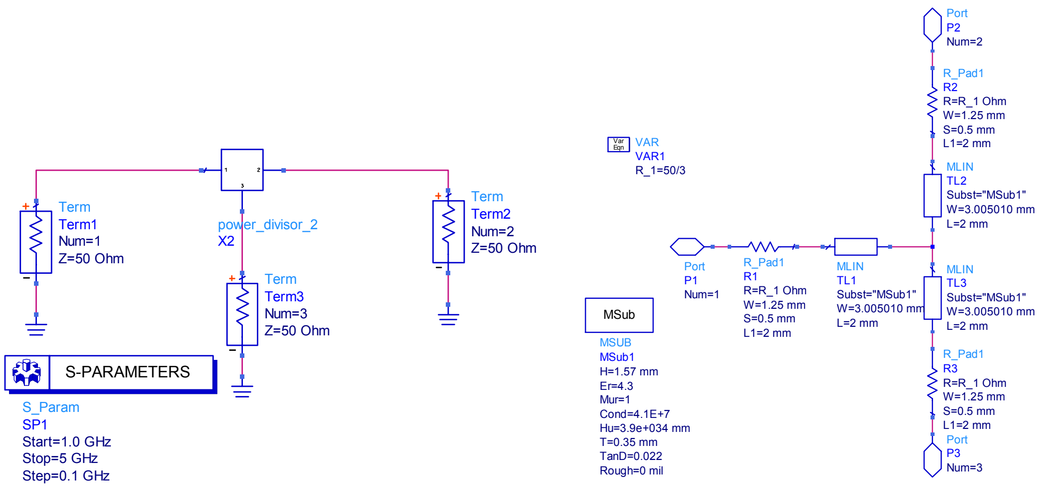


Figure 4.18: schematic of the power divider in ADS

The diagram located to the right of Fig 4.18 is the power divider strictly speaking and to the left is the schematic configuration used to simulate the S-parameters of the net. The results of the simulation can be seen in Fig 4.19.

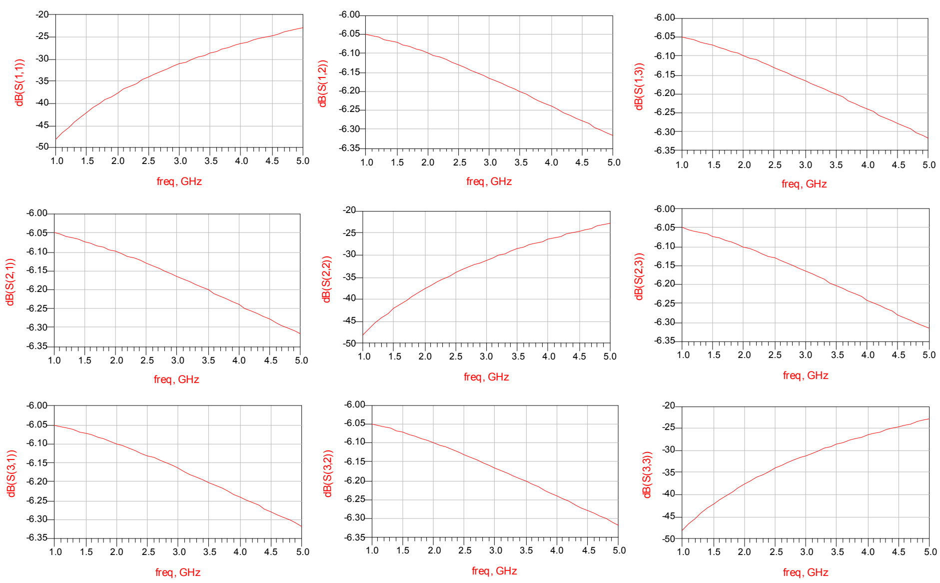


Figure 4.19: S-parameters of the power divider

As expected, if S_{11}, S_{22}, and S_{33} are considered, the ports match is achieved. Approximately 6dB losses from the input to each output can be observed from the other S-parameters. As explained before, these 6dB losses are the 3 dB losses due to the resistive elements plus 3 dB losses because the power is divided in two paths. In conclusion, the theoretical operation of the resistive power divider has been proved by means of its simulation results.

Due to this component fulfils the requirements desired for the purposes of this work and its simple mode of operation, a resistive power divider has been used for the receiver.

v. Mixer

At this point, all components work at radio frequency. The radio frequency supposes the most difficult stage in the design of a transceiver. Some of the difficulties that can be found are the quick phase shift or the coupling between two close transmission lines. Therefore only few of its components should operate at RF.

The function of the mixer is down-converting the received signal at radio frequency to an intermediate frequency much lower by mixing this signal with another signal (the local oscillator (LO)). From the mixer, all components work at much lower frequency, avoiding the undesirable effects occurring when working at radio frequency.

When choosing the mixer, several factors must be taken into account. Maybe the most important is the isolation of the ports. Due to the isolation between the LO and RF ports of the mixer is not perfect, a finite amount of feedthrough exists from the LO port to other points in the circuit. This effect was explained in the chapter 3.a and is called "LO leakage". This effect is very important when a zero IF architecture is used, due to the DC offsets generated. When low IF architecture is used, the problem of DC offset is not so important but another effect is produced. This effect is captured in the following diagram:

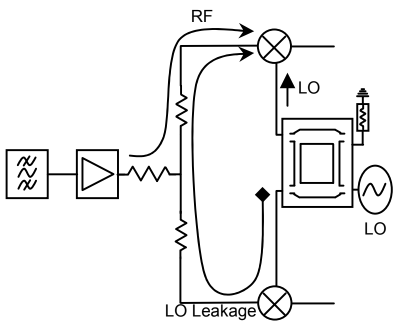


Figure 4.20: Effect of the LO leakage

The leakage signal appearing at the RF input of the mixer goes through the resistor of the power divider and appears at the RF input of the other mixer disturbing its operation. In conclusion, the quality of the IF signal at the output of the mixer is decreased. This problem becomes even more important when a high level local oscillator is used. For this reason, a high isolation mixer must be used.

Another important factor in the mixer is its IP3. The mixer is the component that contributes in a greater extent to this parameter. As mentioned in previous chapters, the

IP3 can limit the dynamic range of the system. Therefore the IP3 requirements of the system have to be observed when choosing the mixer.

Therefore the most important parameters of the mixer are its LO-RF isolation and its IP3. In this section, taking into account these parameters, the mixer is chosen to fulfil the receiver requirements.

The mixers have also a limited range of operation frequencies. Each input can operate only with determined frequencies. The mixer has to be chosen with a range of operation frequencies in each input which includes the frequency range of the desired signals in each input.

The model chosen and its more important features are shown below.

The mixer chosen is the monolithic surface mount mixer model ADE-3G+. Concretely two mixer are used to down-convert the RF signal in both paths of the circuit.

Figure 4.21: Mixer model ADE 3G

Its main features are summarized in the next table:

Model Number	RF in @1dB Comp.	Frequency Range (MHz)		Conversion Loss (dB)			LO-RF Isolation	LO-IF Isolation	IP3@center band
	(dBm)	LO/RF	IF	Typ.	σ	Max.	(dB)	(dB)	(dBm)
ADE-3G(+)	1.0	2300/2700	DC-400	5.6	0.1	7.0	36	26	13

Figure 4.22: Features of the mixer model ADE 3G

As can be read from the table, this mixer fulfils all the required features. Some of these features are compared with the minimum requirements in the next lines.

Its frequency range of operation covers from 2.3 to 2.7 GHz in the LO and RF inputs. Both the frequency of the LO (2.4497 GHz) and the RF signal (2.45 GHz) are covered for this range. The RF signal is down-converted to 300 kHz (IF). Therefore, the frequency range of the IF output of the mixer is enough.

The compression point of the mixer (1 dBm) is quite high. Therefore the power at this point of the circuit is out of the expected power levels.

Taking into account that the mixer is the component which contributes in a greater extent to the IP3 of the receiver circuit, the IP3 of this component (13 dBm) gives enough margin when compared with the minimum IP3 of the receiver front-end (-5 dBm).

As can be seen in table …, the mixer chosen has very high LO-RF isolation (36 dB). Values higher than this are difficult to find in a mixer. Therefore this mixer has been chosen for the receiver circuit. When a higher isolation is required another techniques can be used. An example given in a previous section is the use of a Wilkinson power divider instead of the resistive power divider.

Taking into account that two amplifiers with approximately 18.5 gains are used in the first stage of the circuit, losses of less than 7 dB in the mixer are acceptable.

In the next lines some details of the mixer given in its data sheet are shown. The reader can find more details in this document included in the bibliography.

Firstly the port isolation of the mixer versus the LO level is given:

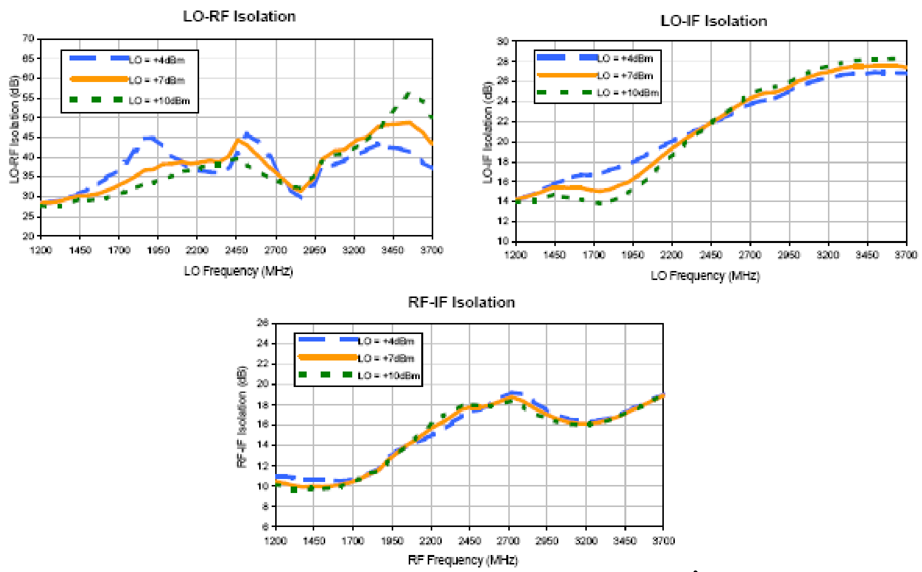
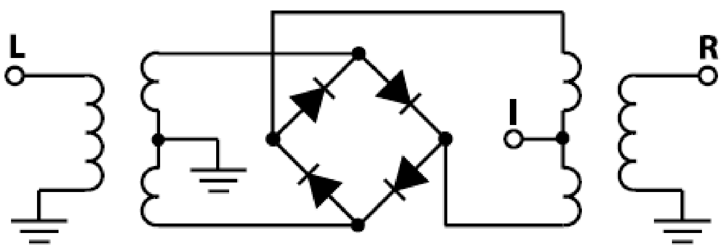


Figure 4.23: Isolation of the mixer versus LO level

The electrical schematic of the mixer is shown in the next diagram:



vi. Polyphase filter

The design of this circuit and the explanation of how the polyphase filter work is the most important step of this work.

The function of the polyphase filter is to reject the image frequency. That is, at the polyphase filter input the desired signal and the image signal are mixed and at the output only the wanted signal occurs.

The image rejection is a difficult issue. The image frequency before the mixer is too close to the desired frequency ($2 \cdot f_{IF}$) because the intermediate frequency is very low. Therefore to reject the image with a preselection filter is very difficult and a very complicated filter must be used. After the mixer, the image frequency is present at the desired frequency band. Therefore a frequency filter can not be used to reject the image.

The problem of the image frequency has been explained in different chapters. A more detailed explanation is given in this section.

The image rejection issue is even more complicated due to the fact that the transmitted signal can not be considered in a frequency point but a range of frequencies. Therefore the rejection of the image can not be done for a frequency point but for a range of frequencies. Due to this aspect, the polyphase filter has to be designed with more than one stage.

In this section a detailed definition and the possible applications of the polyphase filter are given. After that, the signals which arrive to the polyphase filter are deduced from the signals which arrive at the antenna. Sines and cosines signals are introduced to explain easily mathematically the work of the polyphase filter.

After that, the behaviour of the polyphase filter is explained. Once the signal at the input of the polyphase and the operation of the polyphase filter are know, the transformations that these signals suffer from must be clarified. By this way the rejection of the image is demonstrated mathematically.

The frequency of operation of the polyphase filter depends on the values of its components. Therefore how to choose the values of its components is given in this chapter as well. As mentioned above, the transmitted signal can not be considered in a frequency point but a range of frequencies. How to design the polyphase filter to treat with a range of frequencies is explained too. Some simulations are given to check the validity/quality of the mathematical relationships.

At the end of this section, several simulations with the program ADS are shown to clarify graphically this image rejection.

To simulate this component, a schematic block has been created in ADS which includes the net of resistors and capacitors, which make up the polyphase filter, and the baluns used to generate the balanced and unbalanced signals in each case. The baluns will be explained in this chapter. This block allows simulating the polyphase filter as a 4-port network. This block can be seen in below:

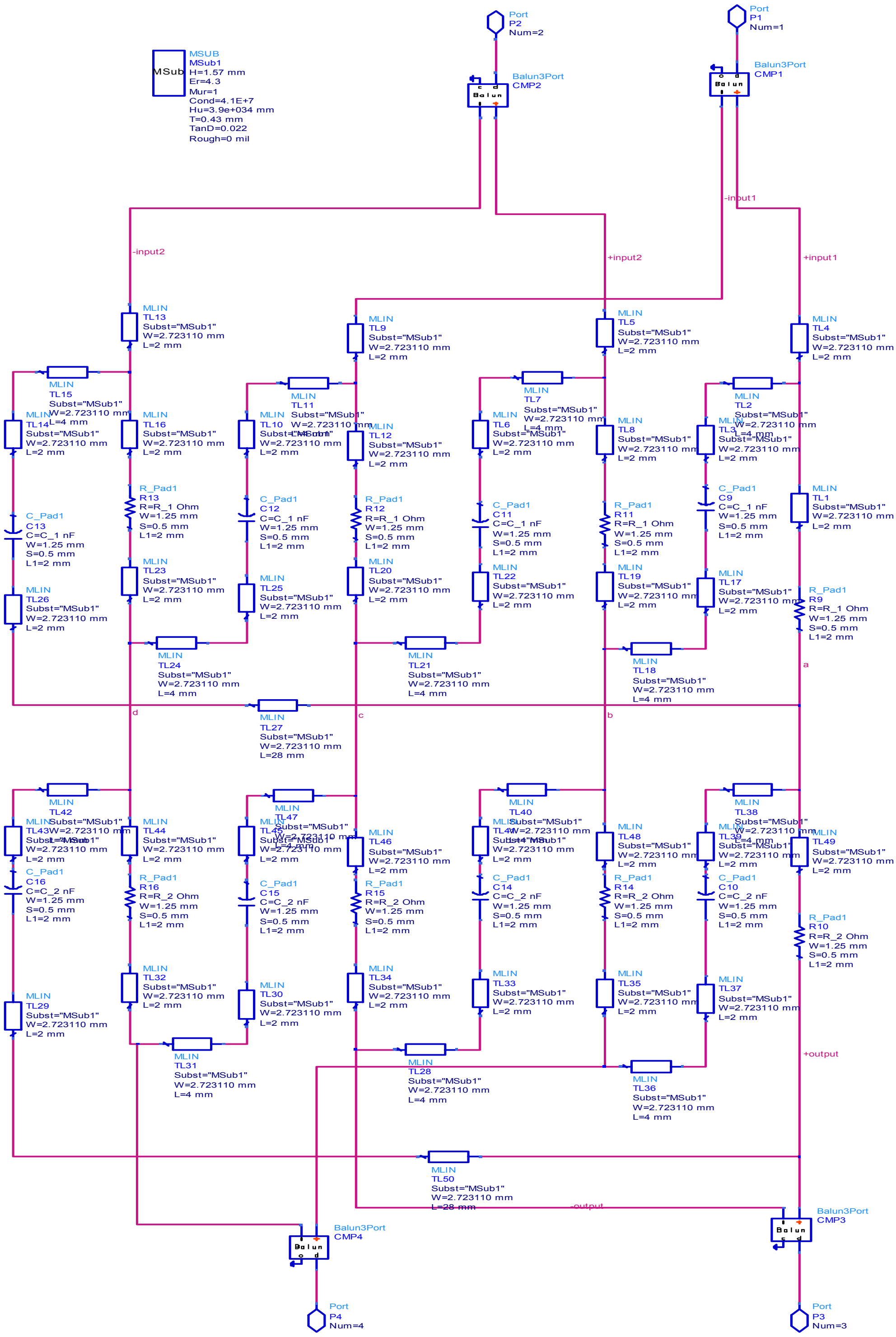


Figure 4.24: Schematic of the Polyphase filter in ADS

The values of the components of the net (RC) are introduced by a component of ADS and can be varied for each simulation.

A polyphase filter is a sequence of asymmetric polyphase networks. A two stage polyphase phase-shifter is shown in Fig 4.24. The RC time constant of each R1 / C1 and R2 / C2 section determines the frequency at which the amplitude and the phase errors are zero. By cascading several sections a broadband phase-shifter can be made. Due to the cyclical connection of the components, a polyphase phase-shifter is less sensitive to component mismatch than an RC − CRs ones. A disadvantage of this structure is the attenuation introduced by the network.

The polyphase filter offers several possibilities for use. It can be used as a 90° polyphase shifter which also introduces only small amplitude errors. The low input impedance of the network dictates the use of an LO buffer amplifier.

But the main application of the polyphase filter is as an IF-Phase shifter to cancel unwanted frequencies.

The IF phase-shifter has to preserve the waveform of the input signal and it requires much wider relative bandwidth. A phase-shifter with these characteristics can be implemented with the help of all-pass filters, but this solution relies on the subtraction of two large signals at the output of the phase-shifter and is very sensitive to amplitude and phase errors. A better solution is to use a polyphase filter like a two stage filter. The key difference with respect to the all-pass phase-shifter is that in a polyphase network the unwanted sideband is cancelled within the network itself and not at the output. This makes the polyphase filter an order of magnitude less sensitive to component mismatch.

For the explanation of the functional principle of the polyphase filter two singletone input signals are assumed for simplicity. In Fig. 4.25 the signals which occur at the input of the polyphase shifter are shown in the frequency domain.

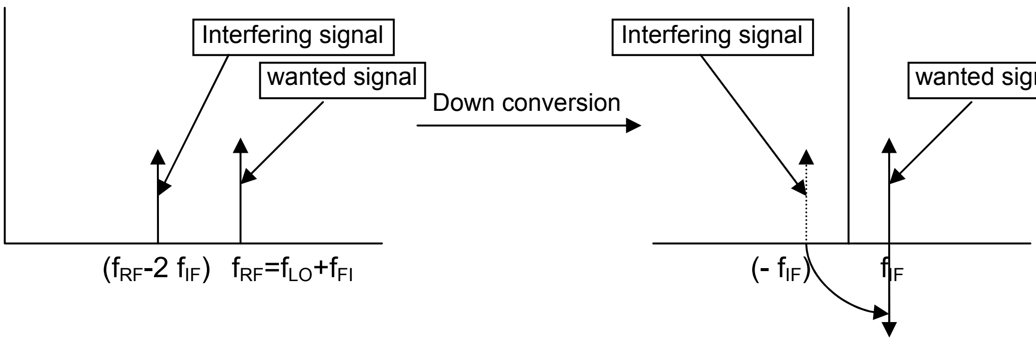


Fig 4.25: problem of the image rejection

At the input of the receiver the following signals which can affect our design are collected:

Wished signal → $S(t) = As \cdot \sin[(\omega_{LO} + \omega_{IF})t]$ (4.34)

Image signal → $M(t) = Am \cdot \sin[(\omega_{LO} - \omega_{IF})t + \Delta\delta]$ (4.34)

The term $\Delta\delta$ in the image signal is included to indicate that the phase of both signals doesn't have to be the same.

The signal at $(\omega_{LO} + \omega_{IF})$ frequency is the interesting signal whereas the other one expresses its image part. This signal is down-converted to an intermediate frequency ω_{IF} by the receiver circuit. The problem is that when the signals at $(\omega_{LO} \pm \omega_{IF})$ are down-converted, both signals are placed at ω_{IF}. Therefore, to reject the signal at $(\omega_{LO} - \omega_{IF})$ is the purpose of the polyphase filter.

As explained in the chapter "Proposed Design" a low-IF architecture is used in this work. This architecture down-converts the signal to ω_{IF} and obtains the signal in phase (path I) and quadrature (path Q) using a local oscillator and a coupler to shift the phase of the local oscillator by 90°.

Supposing that the local oscillator is $G \cdot \sin(\omega_{LO} \cdot t)$, the signal in path I is multiplied by the local oscillator ($G \cdot \sin(\omega_{LO} \cdot t)$) and in path Q is multiplied by the local oscillator plus a shift phase of 90° ($G \cdot \cos(\omega_{LO} \cdot t)$).

After this step, by using the trigonometric relationships, the following signals are obtained:

Path I:

$$\frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta) \tag{4.35}$$

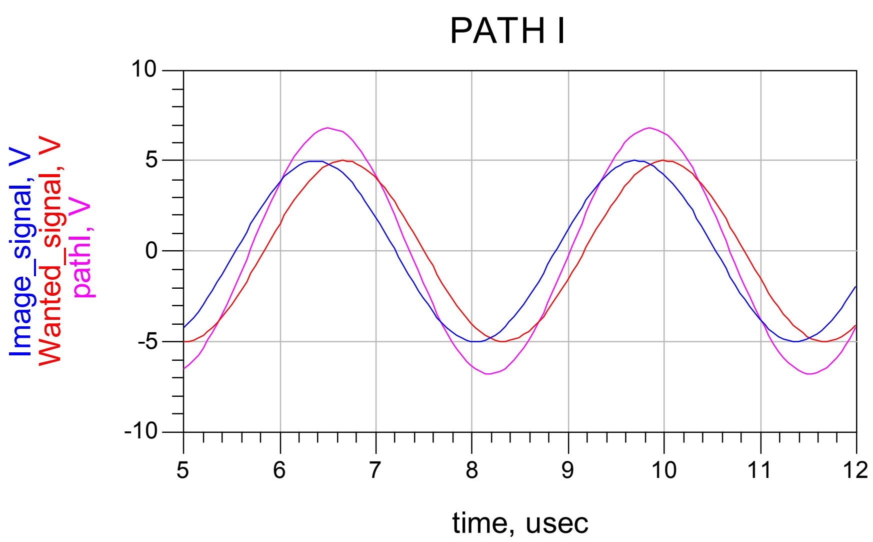


Figure 4.26: Signals in the path I

Path Q:

$$\frac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)-\frac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta) \tag{4.36}$$

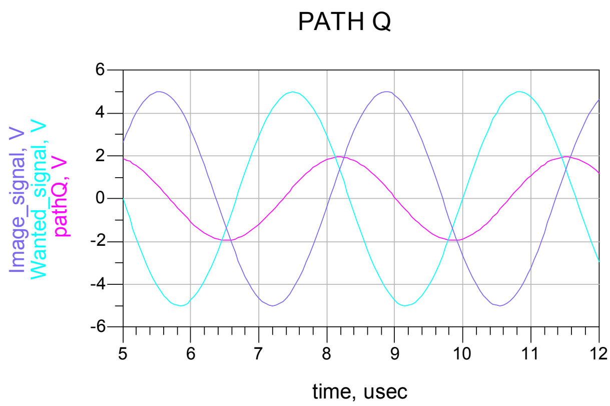


Figure 4.26: Signals in the path I

Now the objective is to reject the signals:

$$\text{Image signals}\begin{cases}\dfrac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta)\\ -\dfrac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta)\end{cases} \tag{4.37}$$

Differential signals at the input of the polyphase filter are needed. Therefore a differential amplifier is used.

Differential signal (path I)

$$\begin{cases}\dfrac{G}{2}\cdot As\cdot\cos(\omega_{IF}\cdot t)+\dfrac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta)\\ -\dfrac{G}{2}\cdot As\cdot\cos(\omega_{IF}\cdot t)-\dfrac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta)\end{cases} \tag{4.38}$$

Differential signal (path Q)

$$\begin{cases}\dfrac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)-\dfrac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta)\\ -\dfrac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t)+\dfrac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta)\end{cases} \tag{4.39}$$

At this point, the signals occurring at the polyphase filter input are known.

First, a 2-stages polyphase filter is used. To understand how it works, the easiest way is to tune both stages to the same frequency. The description of its operation is based on the circuit in figure … given at the beginning of the chapter.

If a signal is introduced in port1, a phase shift of -45 degrees in each stage of the first path of the polyphase filter ("a" in the picture) is obtained.

If a signal in port2 is introduced, a phase shift of -135 degrees in each stage of the same path mentioned above is obtained. For the second path ("b"), third path ("c") and fourth path ("d") the phase shifts are (+45, -45), (-225, +45) and (-135, -225) repectively.

As can be seen, if two signals are introduced within the inputs of the circuit (input1 = port1; input2 = port2), the phase shift in the first signal is 90° higher than the phase shift in the second signal, irrespective of the values of the phases of the inputs.

If the inputs of the polyphase filter are considered, it can be noticed that if the signal in the Q path is introduced at input1 and the signal of the I path is introduced at input2, the image is rejected and the wanted signals add in phase. This can be seen mathematically in the next expressions which represent the signals after the first stage (after the phase shift):

$$\text{POINT “a”} \begin{cases} \frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t - 135) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta - 135) + \\ + \frac{G}{2} \cdot As \cdot \sin(\omega_{IF} \cdot t - 45) - \frac{G}{2} \cdot Am \cdot \sin(\omega_{IF} \cdot t + \Delta\delta - 45) \\ = G \cdot As \cdot \cos(\omega_{IF} \cdot t - 135) \end{cases} \tag{4.40}$$

$$\text{POINT “b”} \begin{cases} \frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t - 45) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta - 45) + \\ + \frac{G}{2} \cdot As \cdot \sin(\omega_{IF} \cdot t + 45) - \frac{G}{2} \cdot Am \cdot \sin(\omega_{IF} \cdot t + \Delta\delta + 45) \\ = G \cdot As \cdot \cos(\omega_{IF} \cdot t - 45) \end{cases} \tag{4.41}$$

$$\text{POINT “c”} \begin{cases} \frac{G}{2} \cdot As \cdot \cos(\omega_{IF} \cdot t + 45) + \frac{G}{2} \cdot Am \cdot \cos(\omega_{IF} \cdot t + \Delta\delta + 45) + \\ + \frac{G}{2} \cdot As \cdot \sin(\omega_{IF} \cdot t - 225) - \frac{G}{2} \cdot Am \cdot \sin(\omega_{IF} \cdot t + \Delta\delta - 225) \\ = G \cdot As \cdot \cos(\omega_{IF} \cdot t + 45) \end{cases} \tag{4.42}$$

$$\text{POINT "e"}\begin{cases} \frac{G}{2}\cdot As\cdot\cos(\omega_{IF}\cdot t-225)+\frac{G}{2}\cdot Am\cdot\cos(\omega_{IF}\cdot t+\Delta\delta-225)+ \\ +\frac{G}{2}\cdot As\cdot\sin(\omega_{IF}\cdot t-135)-\frac{G}{2}\cdot Am\cdot\sin(\omega_{IF}\cdot t+\Delta\delta-135) \\ =G\cdot As\cdot\cos(\omega_{IF}\cdot t-225) \end{cases} \quad (4.43)$$

The explained behavior strongly depends on the frequency of operation. These frequencies are determined by the time constant of the resistors and the capacitors of the circuit. In the next paragraphs an explanation is given to know the necessary values for the components of the polyphase filter corresponding to a specific frequency.

As mentioned above, the value of the RC time constant of each section determines the frequency at which the amplitude and the phase errors should be zero. It is desired to get the minimum possible error at the frequency of operation (300 kHz). Therefore the values of R and C have to be adjusted to obtain the following equality:

$$f=\frac{1}{2\cdot\pi\cdot R\cdot C} \quad (4.44)$$

In the practical case, it is desired to reject an image with a spread spectrum in the frequency domain. Therefore a polyphase filter with several stagger/tuned stages has to be used and the values of the R and C components in each stage have to be adjusted to obtain a certain bandwidth. The larger the image rejection desired, or the higher the ratio of maximum to minimum signal frequency, the more cascaded polyphase stages are needed. Because the spectrum of the signal in this work is not so wide the use of a two stage polyphase filter is adequate (more stages would imply too many losses). A good value for the rejection bandwidth of the polyphase filter is the bandwidth of one channel with ± 25% added on as margin. As the bandwidth of the received signal is 300 kHz, a good value for the rejection bandwidth BWPF is:

$$\begin{aligned} &BW_{PF}=300\cdot 10^3+2\cdot 0{,}25\cdot 300kHz=450kHz \\ &\xrightarrow{Central\_frequency=300kHz} Frequency\_range=[75kHz;525kHz] \end{aligned} \quad (4.45)$$

For the centre frequency the values for the elements have to be to:

$$f=\frac{1}{2\cdot\pi\cdot R\cdot C}\Rightarrow R\cdot C=\frac{1}{2\cdot\pi\cdot 300\cdot 10^3}\Rightarrow(R=10\Omega)\Rightarrow C=53{,}051nF \quad (4.46)$$

To check the net when using these values, the next schematic is created using the block shown in Fig 4.24:

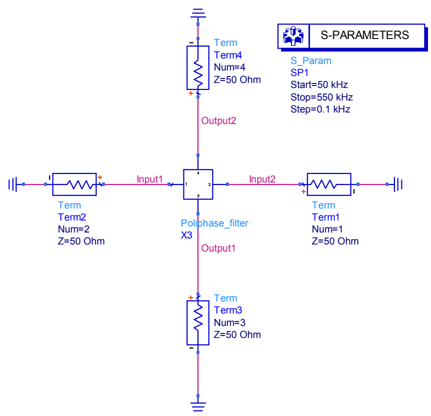


Figure 4.27: Schematic in ADS to measure S-parameters

Introducing the values calculated for R and C, the results obtained are shown below:

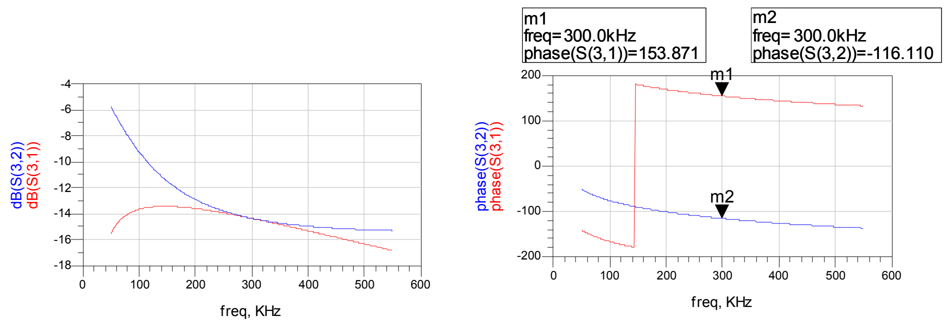


Figure 4.28: Phase and magnitude in dB of the S_{32} and S_{31}

It can be seen from the graphics that the polyphase filter introduces the same losses and 90° phase difference for both inputs at the design frequency (300 kHz). These are exactly the expected results since 90° phase difference between the two input signals introduced in the circuit and the same amplitude shift for both inputs are needed in each stage of the polyphase filter to cancel the image (as explained above). That is, using these values for the resistors and the capacitors, the amplitude and phase errors introduced by the polyphase filter are minimal. Therefore the theoretical calculation for the values of the time constant RC is proven by the simulation.

As mentioned above, the main disadvantage of this structure is the attenuation introduced by the network. Approximately 15 dB losses can be read from Fig 4.28. When introducing the signal in the path I and the signal in the path Q at the same time, the expected losses are less than 15 dB due to the addition of the two signals in phase.

The polyphase filter must not be designed for the central frequency of work, but for a frequency range. This calculated limits are f_1 = 75 kHz (lower limit) and f_2 = 525 kHz (upper limit).

Therefore the values for the components of each stage of the net are:

$$f_1 = \frac{1}{2 \cdot \pi \cdot R \cdot C} \Rightarrow R \cdot C = \frac{1}{2 \cdot \pi \cdot 75 \cdot 10^3} \Rightarrow (R = 10\Omega) \Rightarrow C = 212{,}206nF \quad (4.47)$$

$$f_2 = \frac{1}{2 \cdot \pi \cdot R \cdot C} \Rightarrow R \cdot C = \frac{1}{2 \cdot \pi \cdot 525 \cdot 10^3} \Rightarrow (R = 10\Omega) \Rightarrow C = 30{,}315nF \quad (4.48)$$

Using these values for the net the response of the polyphase filter is:

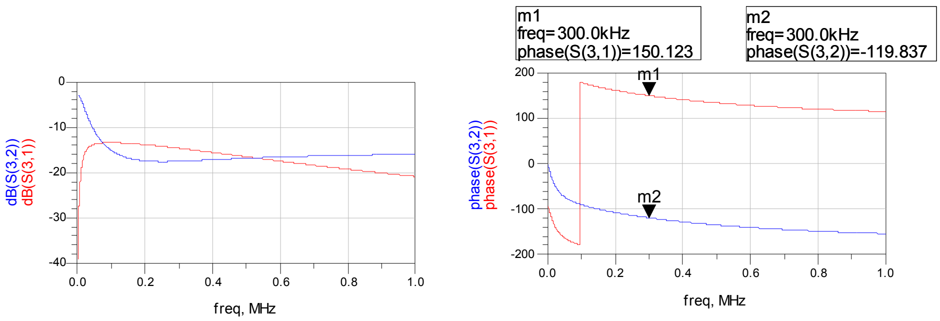


Figure 4.29: Phase and magnitude in dB of the S_{32} and S_{31}

From the diagrams it is obvious that there is a little amplitude error at the centre frequency but a bigger bandwidth with a little amplitude error is achieved (~0.75 MHz).

The next explanation tries to show the time domain behavior of the polyphase filter in the presence of interference at the image frequency. For this purpose, a transient analysis of the whole circuit has been done. To implement it in ADS, four sources for generating the wanted and image signals, two splitters to add the signals to each path and the block shown in figure 4.24 are used. This can be seen in the figure below:

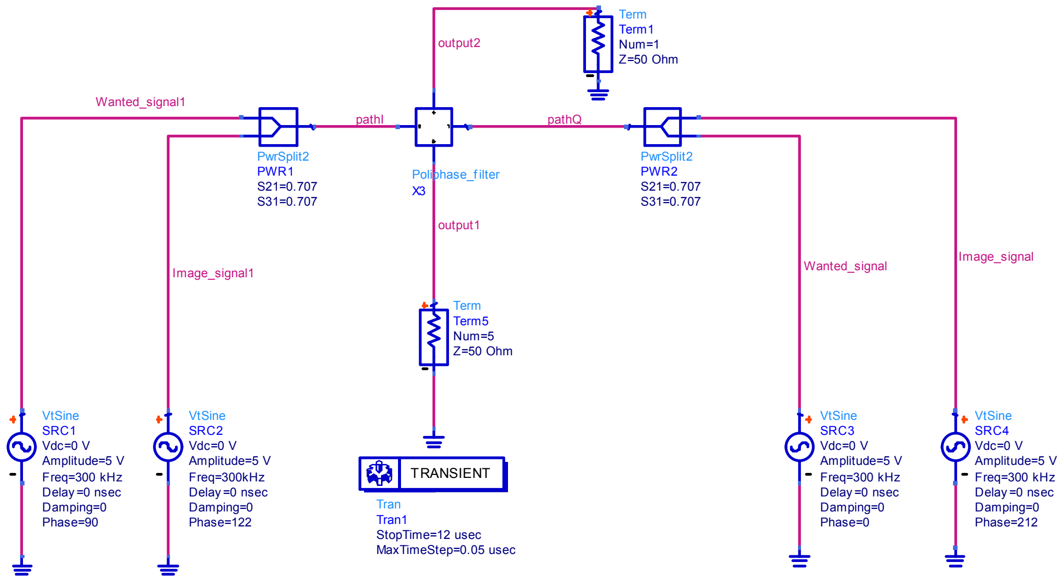


Figure 4.30: Schematic to measure the polyphase filter

The simulation is done by three steps.

The first step consists of setting the sources which generate the image signal to zero volts, in other words, the wanted signal is the only signal introduced into the circuit. In the next diagrams the output signal and the signals introduced in the system are shown:

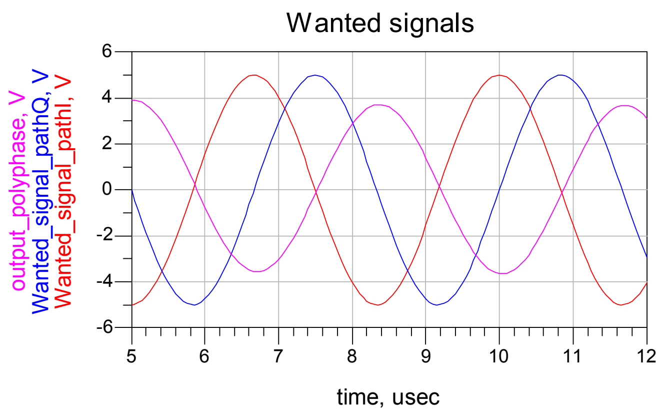


Figure 4.31: Wanted signals

As can be seen, some losses are introduced by the polyphase filter, but the output signal keeps an acceptable level.

The second step consists of introducing the image signal into the circuit by simultaneously nulling the sources of the wanted signal. In the next diagrams the output signal and the input signals introduced of the system are shown:

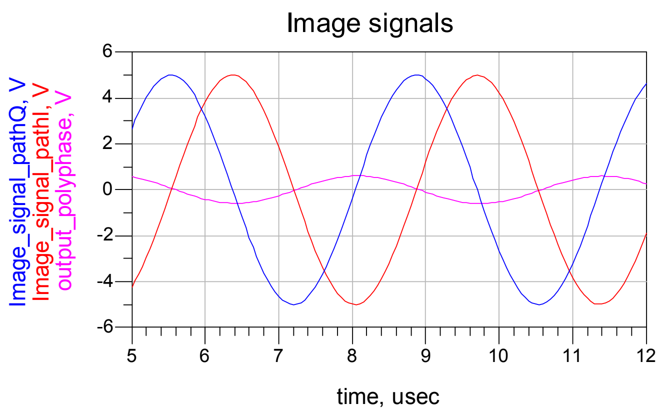


Figure 4.32: image signals

As can be seen, the output level has been reduced considerably. The image rejection is not perfect because the polyphase filter is designed for a certain frequency range. As explained in this chapter, the design of the polyphase filter for a specific band introduces a little amplitude error at the centre frequency which is necessary to get a larger width bandwidth.

Finally, both signals are included at the same time:

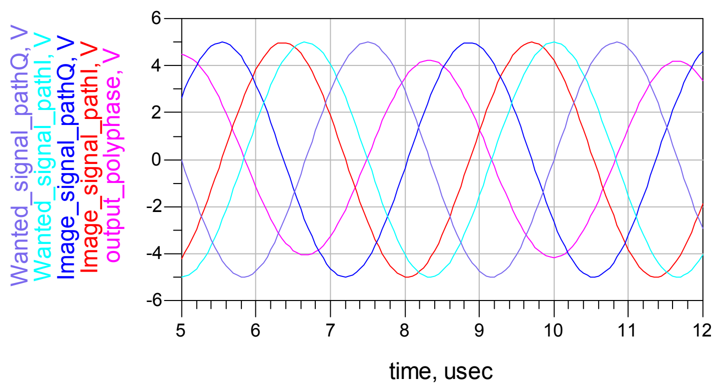


Figure 4.33 output polyphase filter

In this case, the output is very similar to the case in which only the wanted signal was introduced. This confirms that the main contribution to the output comes from the wanted signal.

vii Branch-line Coupler

The signal received by the antenna is amplified and split into two branches. One branch must carry the in-phase component (the component that is in phase with the original carrier) and the other branch must carry the quadrature component (the component that is 90° out of phase with respect to the original carrier).

These two components of the signal are generated at the same time by mixing the signal with the two quadrature signals generated by local oscillators. As explained in a previous chapter, the phase shift and the amplitude shift between the two signals must be 90° and zero respectively. A possibility is to use only a local oscillator to generate the signal at the desired frequency and to split it in two different paths where each signal has the desired characteristics. This shows that a delay must be added to one of the split signals such that the phase shift between the two signals is 90o.

To obtain the desired characteristics in both signals a branch-line coupler is used. A description of the coupler and the main steps in the design of this component are shown in the next lines. A simulation of the S-parameters of an ideal coupler is done to check the quality of the calculated dimensions.

A picture of the geometry of a branch-line coupler and of the geometry of a micro strip line is shown in figure 4.34:

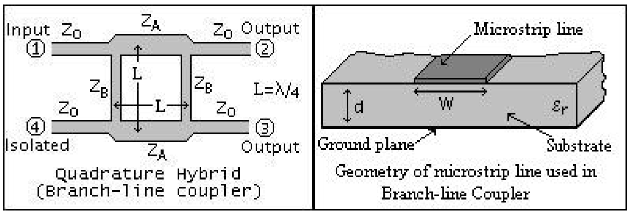


Figure 4.34: geometry of the branch-line coupler

Generally branch-line couplers are 3dB, four ports directional couplers having a 90° phase difference between its two output ports. Branch-line couplers (also known as Quadrature Hybrids) are often made in microstrip or stripline form.

A branch-line coupler is made by two main transmission lines (series line) connected by two secondary lines (branch lines). As can be seen from the figure, it is a symmetrical four port. First port is named as input port, second and third ports are output ports and the fourth port is the isolated port. The second port is also named as direct or through port and the third port is named as coupled port. It is obvious that due to the symmetry of the coupler any of these ports can be used as the input port but at that time the output ports and isolated port changes accordingly.

When designing a coupler, two dimensions have to be paid attention to. Considering the dimension of the coupler the length of the branch line and series line (L) is chosen as a quarter of the design wavelength at the frequency of operation. However, 3/4, 5/4 or 7/4 wavelengths (etc.) could also be used on each arm, the penalty is paid in decreasing bandwidth. Therefore the dimension of the coupler can be expressed as a function of the frequency of operation and the substrate by the next expression:

$$L = \frac{\lambda}{4} \Rightarrow \lambda = \frac{v_p}{f} = \frac{c/\sqrt{\varepsilon_r}}{f} \Rightarrow L = \frac{c}{4 \cdot f \cdot \sqrt{\varepsilon_r}} \tag{4.49}$$

Where εr is the relative dielectric constant of the substrate and f is the operational frequency.

Considering the dimension of the coupler the width of the branch line (W1) and the series line (W2) are chosen to obtain an impedance ZB equal to the characteristic impedance of the line Z0 and an impedance ZA equal to $Z_0 / \sqrt{2}$.

Taking into account the characteristics of the substrate used given in the second section of this chapter and the necessary frequency of the local oscillator (fLO = 2.45 GHz – 300 kHz = 2.4497 GHz) to down-convert the RF signal (fRF = 2.45 GHz) to IF signal (fIF = 300 kHz), L, W1 and W2 are calculated as follows:

$$L = \frac{c}{4 \cdot f \cdot \sqrt{\frac{\varepsilon_r + 1}{2}}} = \frac{3 \cdot 10^8 m/s}{4 \cdot 2.4497 GHz \cdot \sqrt{\frac{4.3+1}{2}}} = 18.8mm \tag{4.50}$$

To calculate the values of W_1 and W_2 the tool Linecalc of the program ADS is used. The parameters introduced into this program are shown in the next picture:

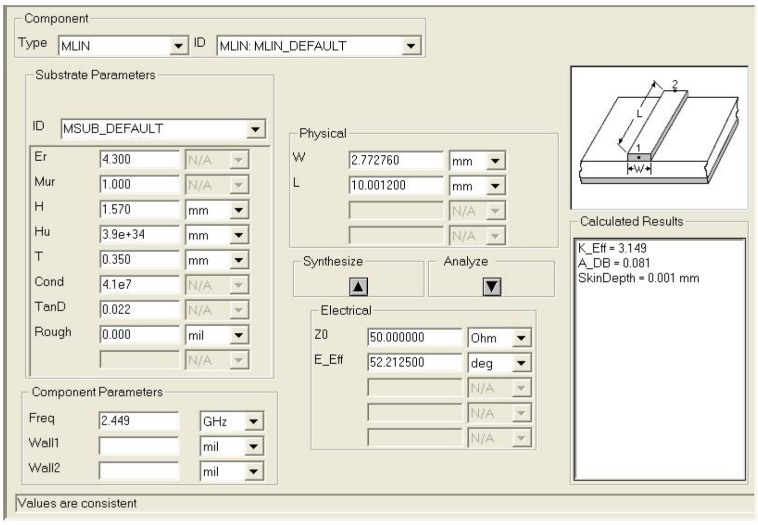


Figure 4.35: Calculation of W_1 by using linecalc

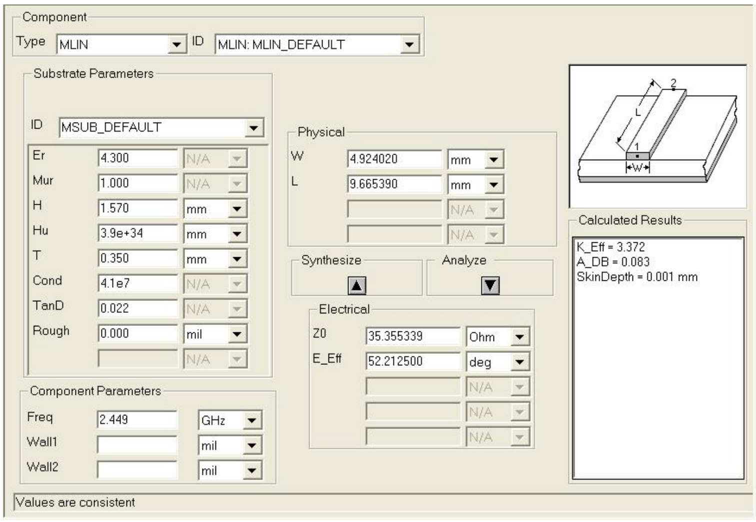


Figure 4.36: Calculation of W_1 by using linecalc

With respect to the fixed parameters of the layout (see paragraph about impedance of micro strip line) the width of the lines are W2= 4.924020 mm and W1 = 2.7727 mm.

A simulation of the S-parameters the coupler with the calculated dimensions can be seen in the next figures.

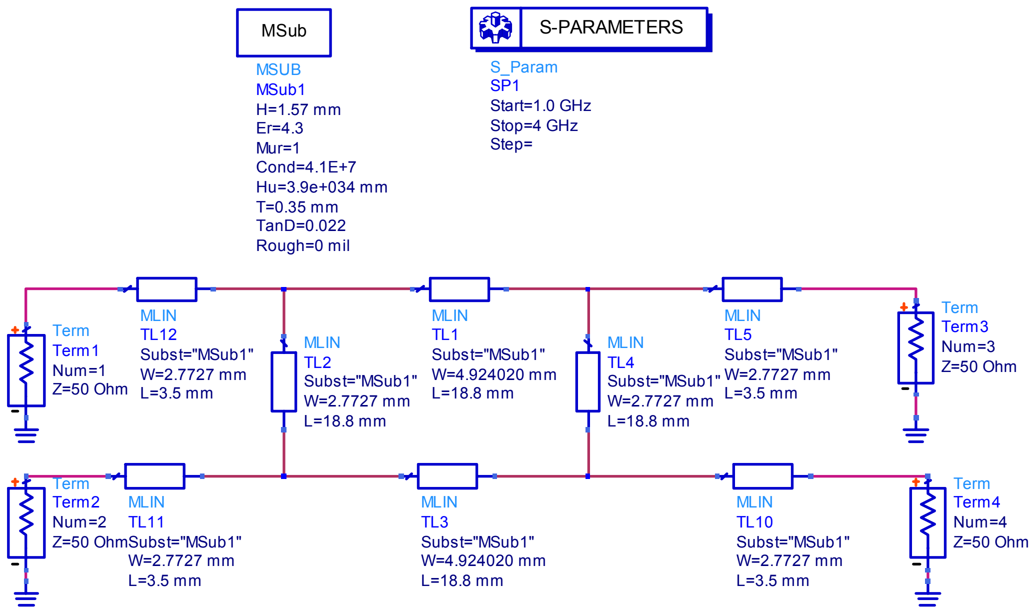


Figure 4.37: Schematic of the branch line coupler

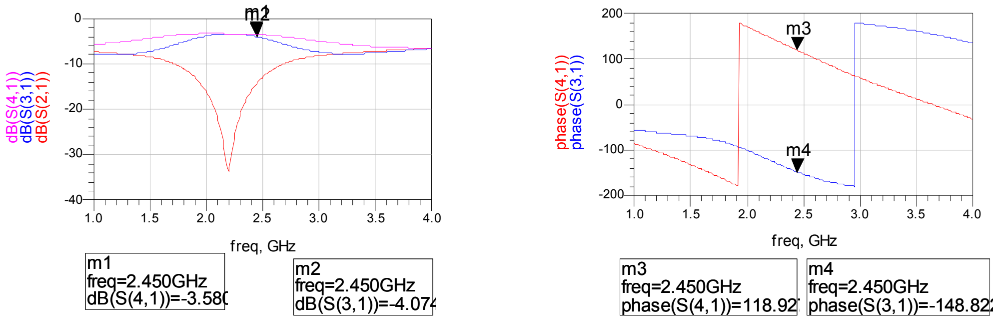


Figure 4.38: Phase and magnitude in dB of the S_{41} and S_{31}

The graphs reveal that for the calculated values calculated the amplitude and phase error deviate from zero for the frequency of work. Concretely 0.494 dB amplitude error and 2.251° deviation from the desired phase shift can be mentioned.

By tuning the parameters of the simulation, the length of the lines which achieve zero amplitude and phase errors is 16.9 mm. The graphs obtained by using these new values are shown below.

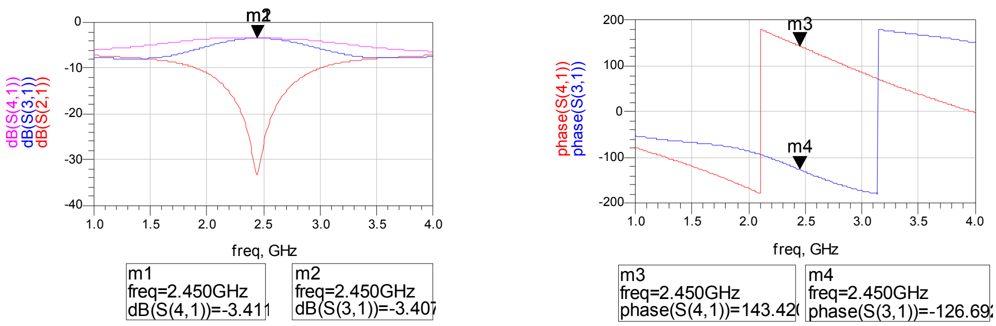


Figure 4.39: Phase and magnitude in dB of the S_{41} and S_{31}

The value used in this work for the length of the transmission lines is 18 mm. By means of real measurements, the best performance of the coupler has been obtained by using these dimensions of the lines.

viii. Baluns

As mentioned in a previous chapter, the polyphase filter works with balanced signals but the receiver circuit works with unbalanced signals. Therefore, balanced and unbalanced signals must be generated from each other. The terms “balanced” and “unbalanced” are used in engineering to refer to differential and single-ended transmission.

The generation of balanced signals from a single-ended input is simple: the input signal is inverted and buffered to create two output signals with a phase difference of 180º. A more detailed explanation can be found in the references of this chapter. The problem is that such approximation suffers from different phase and amplitude gains in each output. Although this problem can be overcome by different circuits, an easier solution has been adopted for this work.

In this work, a differential-input and differential-output amplifier is used to generate differential signals from single-ended signals. The main reasons of this choice are its simplicity of application and its high level of achievable performance.

Concretely, two of these amplifiers are placed in the circuit to generate the two differential inputs for the polyphase filter. It can be seen in figure…

The chosen model, a general description, its main features and the application circuit are explained in this section. More detailed information can be found in the data sheet of this element.

The model chosen is the ADC driver AD8137. The AD8137 is a low cost differential driver. It is easy to apply and has an internal common-mode feedback architecture which allows its output common-mode voltage to be controlled by the voltage applied to one pin. The internal feedback loop also provides inherently balanced outputs as well as suppression of even-order harmonic distortion products. Fully differential and single-ended-to-differential gain configurations are easily realized by the AD8137. External feedback networks consisting of four resistors determine the amplifier’s closed-loop gain. The power-down feature is beneficial in critical low power applications.

The main features of this device are shown in the next table:

Model Number	-3 dB BW	Min Gain	Voltage Supply	Supply Current	Slew rate	Distortion (2nd)	Distortion (3nd)
	(MHz)	(Acl)	(V)	(mA)	(V/us).	(dBc)	(dBc)
AD8137	110	1	+/- 6	3.6	450	-85	-85

Table 4.4: Features of the differential amplifier model AD137

As can be seen and taking into account that the AD8137 has to handle signals at the intermediate frequency (300 kHz), these features are far above of the requirements of the receiver.

A typical connection for the AD8137, using matched external R_F/R_G networks is shown in Fig 4.40

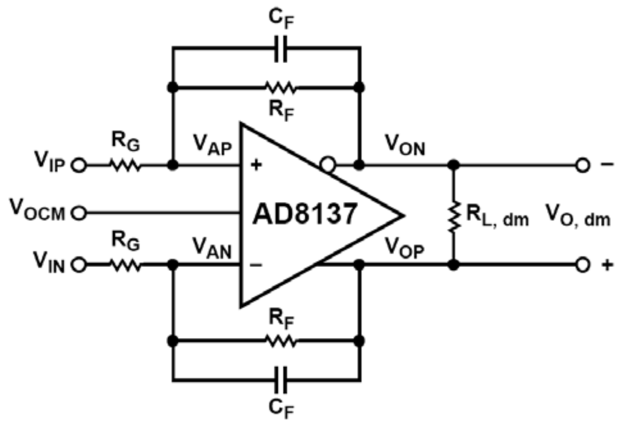


Figure 4.40: typical connection for the AD8137

Depending on the closed-loop gain, different features of the amplifier can be achieved. The next table shows the recommended values of gain-setting resistors and voltage gain for various closed-loop gains.

Gain	RG (Ω)	RF(Ω)	3 dB Bandwidth (MHz)
1	1k	1k	72
2	1k	2k	40
5	1k	5k	12
10	1k	10k	6

Table 4.5: recommended values of gain-setting resistor and voltage gain

The application circuit used in this work can be seen in the next diagram:

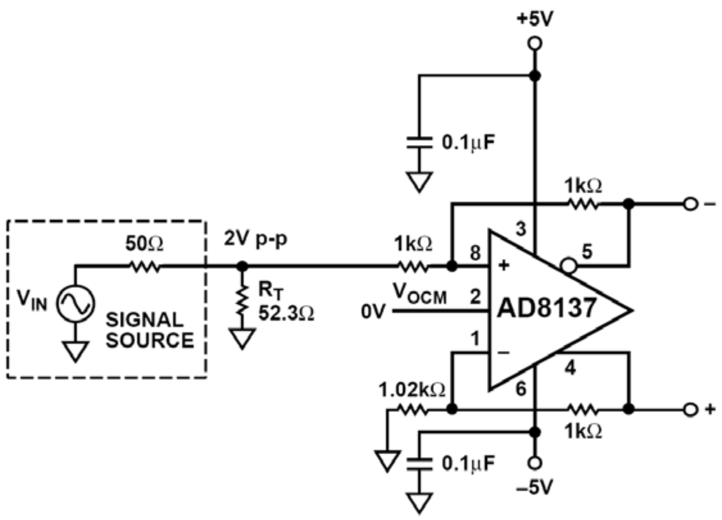


Figure 4.41: application circuit

This configuration is called unity-gain configuration and provides a proper termination in a 50 Ω environment. The 52.3 Ω termination resistor, RT, in parallel with the 1 kΩ input resistance of the AD8137 circuit, yields an overall input resistance of 50 Ω that is seen by the signal source.

In order to have matched feedback loops, each loop must have the same RG if it has the same RF. In the input (upper) loop, RG is equal to the 1 kΩ resistor in series with the positive input plus the parallel combination of RT and the source resistance of 50 Ω. In the upper loop, RG is therefore equal to 1.03 kΩ. The closest standard value is 1.02 kΩ and is used for RG in the lower loop.

Things become more complicated when it comes to determining the feedback resistor values. A detailed calculation of this is given in the data sheet of this component included in the references of this chapter.

For the same reasons explained for the single-to-differential case, a differential to single-ended amplifier is used for generating a single-ended output from balanced signals at the input.

This amplifier is the last stage of the receiver circuit. The function of this amplifier in the receiver circuit is to generate a single-ended signal from one of the two differential outputs of the polyphase filter. It is possible to use two of these amplifiers for both outputs to obtain the two components of the signal (in-phase and quadrature components) at the output of the receiver circuit.

The chosen model, a general description, its main features and the application circuit are explained in this section. More detailed information can be found in the data sheet of this element.

The chosen model is the differential receiver amplifier AD8129. It is designed for the transmission of high speed signals over twisted-pair cables. Either can be used for analog or digital video signals and for high speed data transmission.

The AD8129 has several features interesting for receiver circuits. For example, it is a low noise, high gain (10 or greater) version intended for applications where signal attenuation is significant. It has also user-adjustable gain to help compensate for losses in the transmission line. The gain is set by the ratio of two resistor values. Another good feature is its high input impedance on both inputs, regardless of the gain setting. It has even excellent common-mode rejection (70 dB @ 10 MHz), allowing the use of low cost, unshielded twisted-pair cables without fear of corruption by external noise sources or crosstalk. It has also a wide power supply range from single +5 V to ±12 V, allowing wide common-mode and differential-mode voltage ranges while maintaining signal integrity.

Its main features are summarized in the next table:

Model Number	-3 dB BW	Min Gain	Voltage Supply	Supply Current	Slew rate	0.1 dB Flatness
	(MHz)	(Acl)	(V)	(mA)	(V/us).	(MHz)
AD8129	200	10	+4.5V to 25.2V	10.8mA	1060	20

Table 4.6: Features of the differential amplifier model AD129

As well as was seen for the AD8137 these features are far above of the requirements of the receiver.

In the next diagram the typical connection configuration is shown.

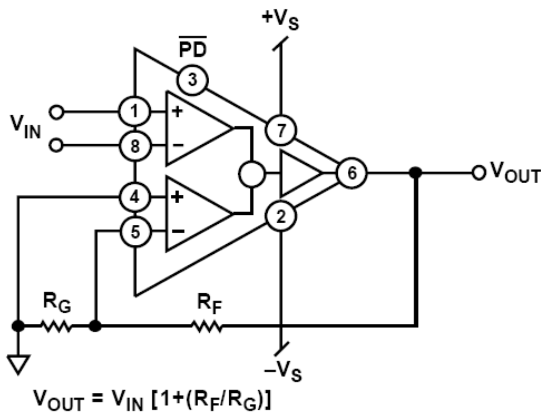


Figure 4.42: typical connection configuration

As shown in the diagram, the gain of the amplifier depends on the ratio of the resistors RG and RF. In this work, the values of the resistor have been chosen to obtain a relation of 10 which results in a linear voltage gain of $G_V = \frac{Vout}{Vin} = 11$.

The reader is referred to the data sheet of this device to know more details of its operation and its different applications.

Chapter 5

Setups and Measurements results

In the previous chapter all components for the receiver chain were chosen and their corresponding characteristics were explained. The whole planned receiver can be seen in Fig. 5.1.

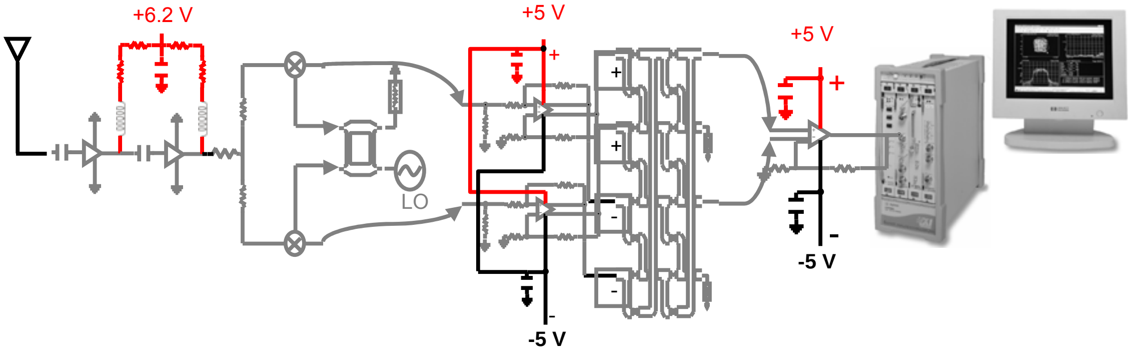


Figure 5.1: Whole planned receiver

In this chapter the measurements of standalone blocks and the receiver chain are carried out. To do this several aspects have to be taken into account. The isolation of the components as well as their dimensions and the available material are some of these concepts.

The isolation of the circuits is a very important issue due to the presence of interferering signals and other effects like coupling. In this work, special cases have been used to isolate the circuits against environmental influences. These cases and the available sizes of these cases can be seen in the next picture.

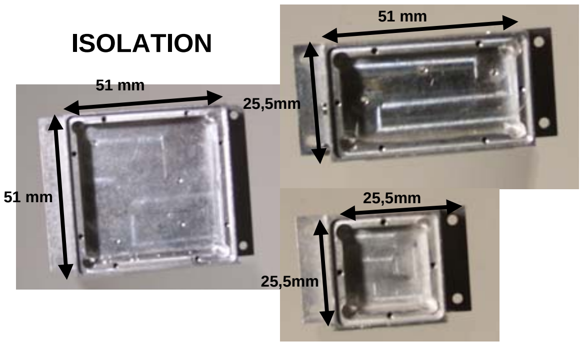


Figure 5.2: Isolation cases and their available sizes

To measure each standalone component of the circuit in a case would imply an excessive cost of time and resources. Therefore, another option has been used. This option consists of dividing the circuit chain in blocks and using a case for each block. The measurement of each block standalone has been carried out by using a test bench which is available in the laboratory and allows validating if the device under test satisfies the properties desired without the mentioned costs.

The test bench used can be seen in the following figure:

Figure 5.3: Polyphase filter in the test bench of the laboratory

Once each component is checked, the choice of the number of cases used and the devices introduced in each case can be carried out.

Due to the dimensions of each component and its application in the receiver chain, the dimensions of the available cases as well as the desired functionality, the complete receiver chain has been divided in four blocks to optimize the available resources. The four blocks are represented in the following figure:

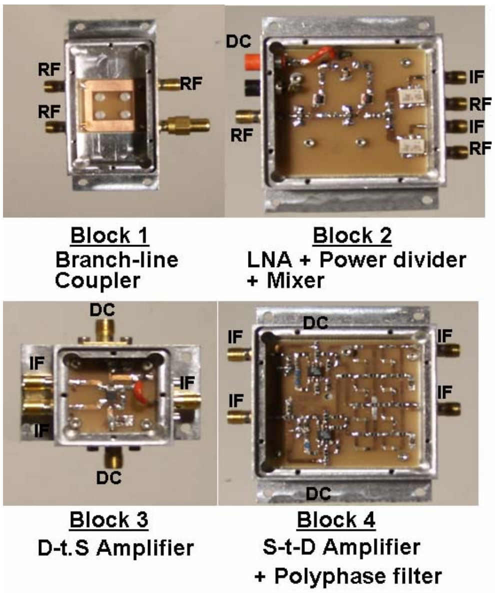


Figure 5.4: Blocks of the receiver

As can be seen from the figure above, the branch line coupler is the only block that has all its ports working at radio frequency (RF). Due to the existence of a VNA available with an operation range from 10 MHz to 60 GHz, the S-parameters of the coupler have been measured. The other blocks have at least one port working at intermediate frequency (IF) (300 kHz). Therefore, other devices with lower frequency of operation have been used to characterize those blocks. The devices used during this work are shown in the following figure:

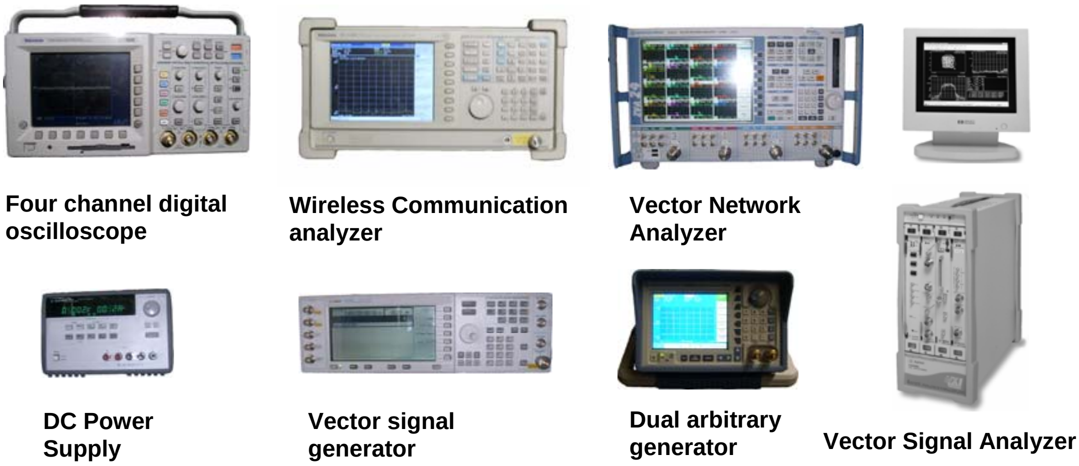


Figure 5.5: Devices used to carry out the measurements

Each setup is described in the next sections. The components which make up the setup, different measurements carried out and an explanation of the achieved results will be given as well. After words, all blocks are combined and the complete receiver is tested. The properties of the receiver chain are checked by using a signal generator to generate the signal normally received by the antenna. An explanation of the obtained results is given as well. Finally the operation of the whole receiver chain is checked in a real environment by measuring the path loss of an indoor channel by employing antennas and transmitters.

a. Block 1

The first block is formed by the branch-line coupler. The branch-line coupler and the four port network analyzer which was used to measure the S-parameters are shown in the next picture.

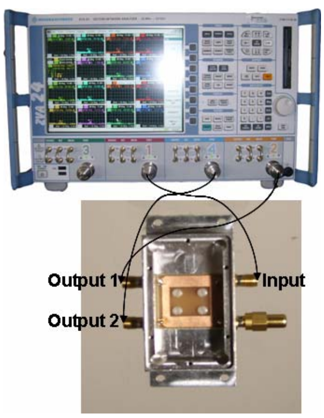


Figure 5.6: Block 1 - setup

As can be seen, the coupler is embedded into a case which is larger than the dimensions of the coupler itself. This is necessary to avoid a short between the signal line and ground which is connected to the case. The plastic screws used for fixing the coupler were chosen to minimize the influence on the behavior of the device.

As described in chapter 4.e.vii, the ideal characteristics of the branch-line coupler are 3dB losses in each output path, port match and 90° phase shift between the two output ports for the operation frequency (2.4497 GHz) as well as one isolated port

To check the port matching, the magnitudes in dB of the obtained S-parameters S_{11}, S_{22}, S_{33} and S_{44} are represented:

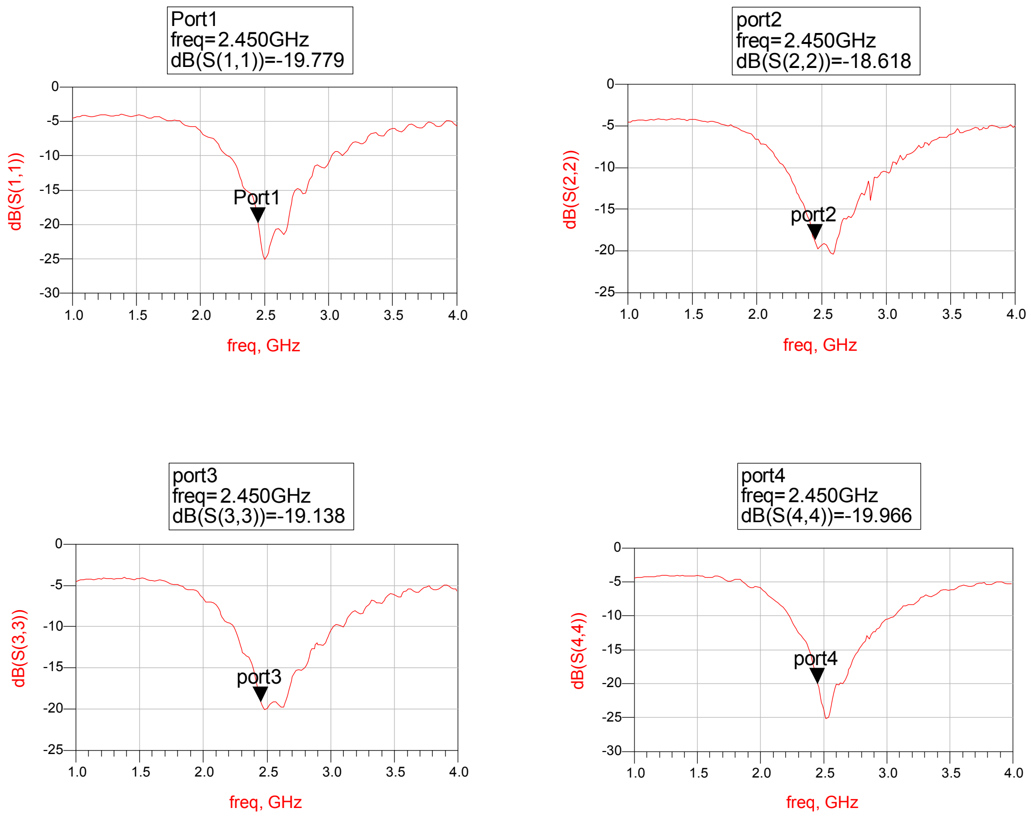


Figure 5.7: Block 1 - S_{11}, S_{22}, S_{33} and S_{44}

As can be seen from the diagrams above, the power reflected from each port is very low, approximately -19.5 dB. Therefore, good port matching is achieved. The curves obtained at all ports indicate that the net is symmetrical as was expected.

To check the losses in each output path and the amplitude mismatch between the output paths, the obtained magnitudes in dB of the S-parameters S_{21}, S_{41} are shown.

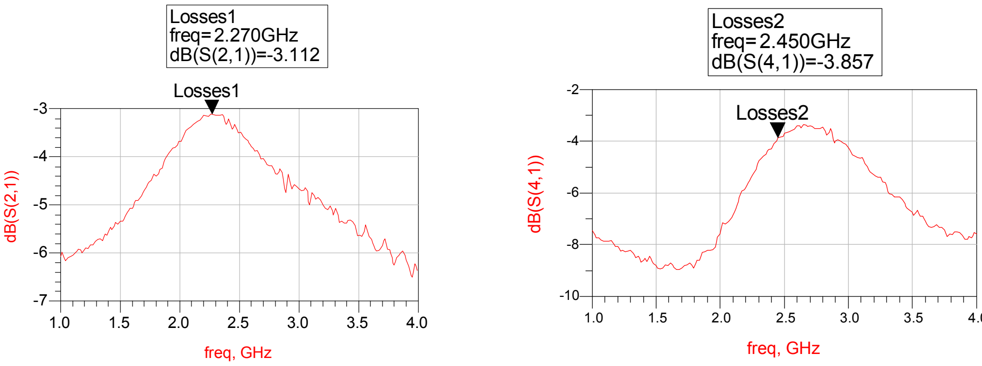


Figure 5.8: Block 1 – dB(S_{21}) and dB(S_{41})

The previous diagrams reveal that, if a signal is introduced into the port 1, the losses are 3.112 dB and 3.857 dB in port 3 and 4 respectively. That is, the measured amplitude error is 0.745 dB. Values lower than this are difficult to achieve. Therefore these occurring results can be considered to be good enough for the receiver chain. However this mismatch superimposes the amplitude mismatch of the receiver channel resulting in a decrease of the IRR.

To check the losses in each path, it is also important to take a look at the phase mismatch

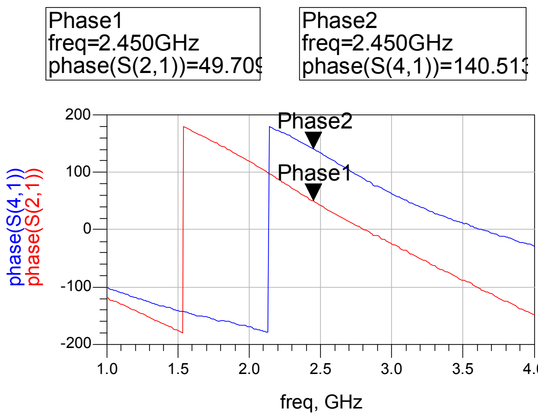


Figure 5.9: Block 1 – phase(S_{21}) and phase(S_{41})

As can be seen, if a signal is introduced to port 1, the phase in port 2 and 4 are 49.706° 140.513° respectively. That is, the measured phase error is 0807°. Values lower than this value are difficulty achieved. Therefore this value is considered to be good enough for the receiver chain. However this mismatch is also added to the phase mismatch of the receiver channel producing a decrease of the IRR.

Concluding, an amplitude and phase mismatch of 0.745 dB and 0,807° are introduced by the coupler. Approximately 3.5 dB losses are introduced from input to each output path.

b. Block 2

The second block contains two low noise amplifiers (LNAs), the power divider and two mixers. Because of the lack of a device to generate two signals at high frequency and with a 90° phase shift the coupler in conjunction with a signal generator was used. Furthermore the amplifiers are active devices. Therefore, a DC supply voltage is necessary.

Interesting to measure is the output response of both signals both outputs (I path and Q path). It is also necessary to measure the LO-IF isolation by feeding a signal to the LO input of a mixer while measuring the signal at the IF output for RF&LO generation.

The first setup used consisted of two signal generators, the branch-line coupler, the block formed by the 2 LNA, the power divider and the 2 mixers, a DC power supply and a wireless communication analyzer. This setup is shown in the next diagram:

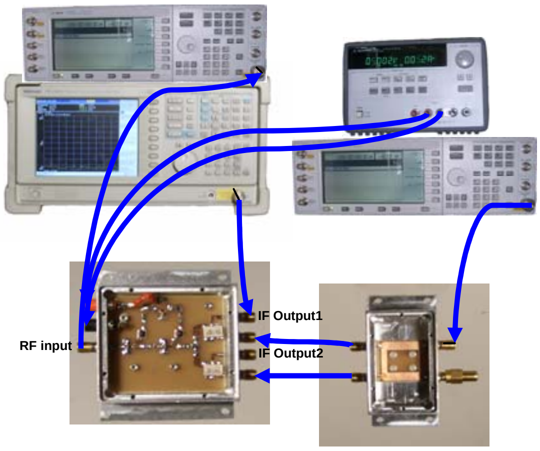


Figure 5.10: Block 2 - Setup

As shown in a previous chapter, the theoretical characteristics of the components of this block are:

	Amplifier (each)	Power divider	Mixer
Gain (dB)	18.4	-6	-7

Table 5.1: Gains of each components of block 2

Therefore the theoretical gain expected for each path is

$$Theoretical\_block\_gain = 2 \cdot 18.4dB - 6dB - 7dB = 23.8 \text{ dB} \tag{5.1}$$

The power spectrum obtained at both IF outputs of the mixers by feeding a -50 dBm signal in the circuit is represented in the following figures:

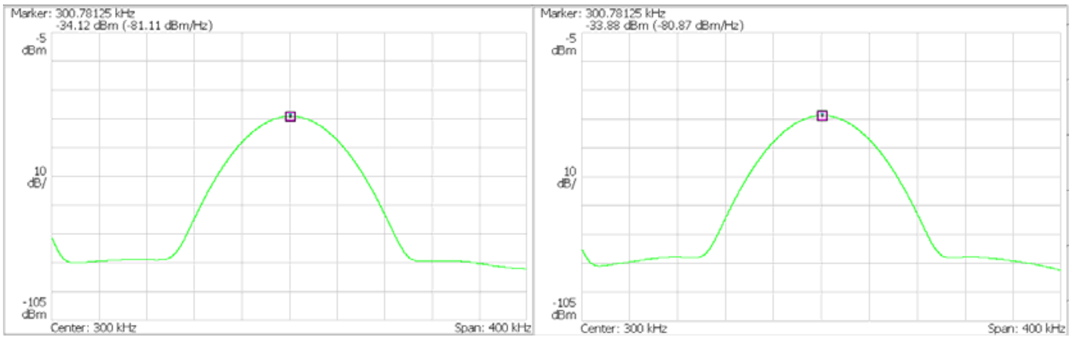


Figure 5.11: Output power spectrums of the two branches of the block 2

Figure : Power spectrum of the signal at the output 1 and output 2

The gain in both paths of the circuit can be deduced from the diagrams taking into account that the input signal power is -50 dBm and the cables used to carry out the measurements introduce 3 dB losses. The Gain at output 1 is:

$$\text{Gain\_output1} = -34.12 - (-3) - (-50) = 18.88\ \text{dB} \tag{5.2}$$

The Gain at output 2 is:

$$\text{Gain\_output2} = -33.88 - (-3) - (-50) = 19.12\ \text{dB} \tag{5.3}$$

Approximately 4.8 dB gain less than the expected are achieved. This can be explained by biasing on layout in accuracy as well as interaction between the different building blocks. Furthermore the expected values were derived from the data sheets base on standalone measurements

The amplitude error measured is:

$$\text{Amplitude error} = 19.12 - 18.88 = 1.76\ \text{dB} \tag{5.4}$$

This mismatch also results in worse IRR.

To measure the phase error, the Wireless Communication analyzer is substituted by a four channel digital oscilloscope to observe the signal in the time domain. The signals at the two outputs of the circuit are shown in the next figure:

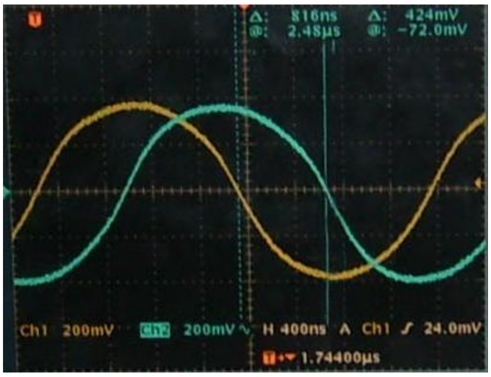


Figure 5.12: Phase shift between the two outputs

The diagram shows an approximately 90° delay in the output 1 related to the output 2. Due to the non-availability devices in the laboratory to measure more accurately the phase shift introduced by the net, the phase error is obtained with this device by using markers. The next picture shows a delay of the output 2 of 816 ns when compared with the output 1. As 3,3 µs correspond with 360°, the delay of the output 2 correspond with:

$$delay = \frac{0.816 \cdot 10^{-6} \cdot 360}{3,3 \cdot 10^{-6}} = 89° \tag{5.5}$$

Therefore, the phase error by using the digital oscilloscope is approximately 1°.

The second setup used consisted of a signal generator, block 2 and wireless communication analyzer. This setup is shown in the next diagram:

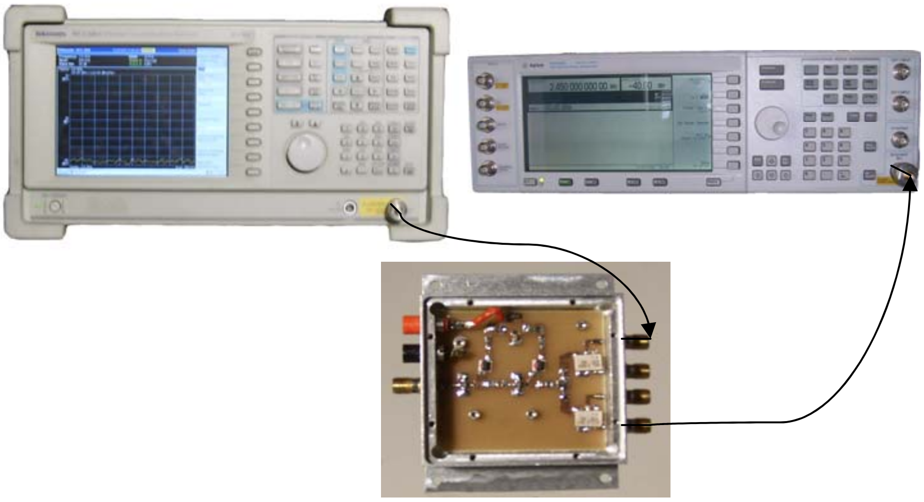
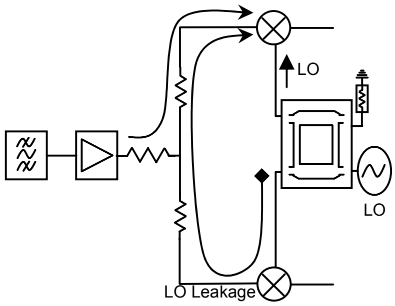


Figure 5.13: Block 2 - setup

As mentioned above, this setup tries to measure the isolation between the LO input of one mixer and the IF output of the other mixer. This effect is represented in the next figure:



The power measured at the IF output of the mixer above of the circuit by introducing 7 dBm signal power into the LO input of the second mixer are shown in the next figure:

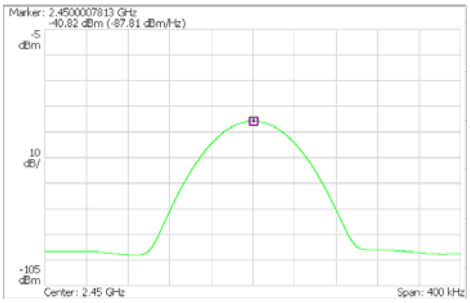


Figure 5.14: LO-IF Isolation

The isolation between both ports is:

$$Isolation = 7 - (-40.82) = 47.82dB \tag{5.6}$$

As can be seen, the isolation is not very good because the LO leakage in the RF input of the mixer above is very high compared with the possible power level of the RF input at this point of the circuit.

c. Block 3

This block contains the differential to single ended amplifier and its application circuit.

The differential to single ended amplifier is the last component of the receiver circuit and provides the circuit with a high gain and a single ended signal output. Therefore, its gain and the quality of the achieved output signal are checked in this section.

The configuration of the measurement setup used can be seen in the next figure:

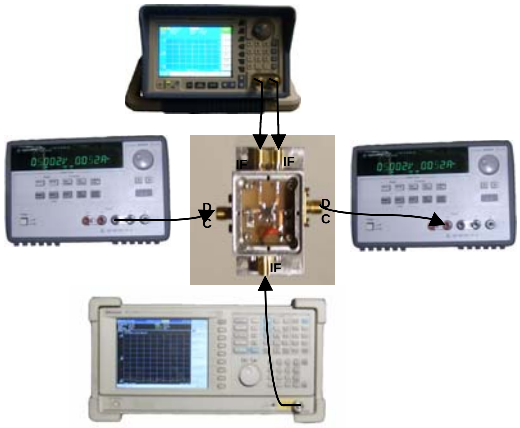


Figure 5.15: Block 3 - setup

The setup used is formed by a dual arbitrary generator to generate two signals with the same amplitude and frequency and 180° phase shift (achieving a differential signal), the block formed by the differential to single ended amplifier and its application circuit (two DC power supplies) and a wireless communication analyzer.

This setup gain and functionality of the block.

The power spectrum obtained by using this setup and an -50 dBm input signal power is shown in the following figure.

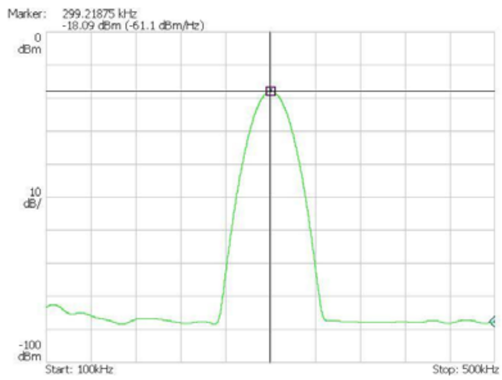


Figure 5.16: Output power spectrum block 3

-18.09 dBm output power can be seen in the previous figure. Therefore the gain of this circuit is -18.09 – (-50) = 31.91 dB.

From the clear spectrum achieved, a good quality of the output signal can be deduced.

Block 4

Block 4 contains two single ended to differential amplifier and a polyphase filter. Here block 3 has been used to carry out the measurements because the measurement devices only support single ended signals.

The main property of this block is the image rejection. The losses introduced by this block in the received signal are checked in this section as well as the imbalance of the outputs and the image rejection. The configuration of the measurement setup used can be seen in the next figure:

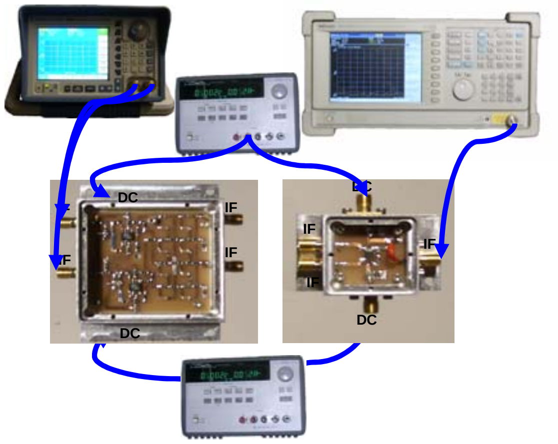


Figure 5.17: Block 4: setup

This setup consists of a dual arbitrary generator to generatethe needed I-Q input signal normally provided by the mixer output, the block formed by 2 single to differential ended amplifiers and the polyphase filter, furthermore DC sources for biasing and a wireless communication analyzer are used.

The losses introduced by this block will be checked with this setup. This setup also allows determining the image rejection by changing the phase relation of the input signals as explained in chapter 4.3.

As mentioned in a previous chapter, the IRR as a function of the amplitude and phase mismatch can be seen in the next diagram:

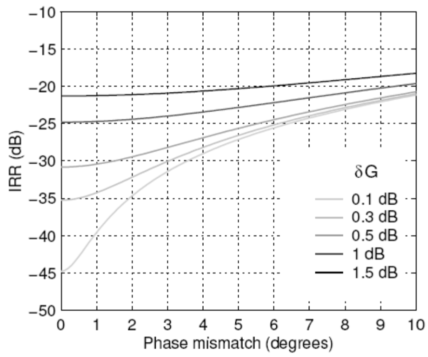


Figure 5.18: IRR

The results obtained by using this setup are shown. Three different results are shown. In the first case, the in-phase and quadrature components of the image signal are generated by the dual arbitrary generator. -50 dBm signal power was fed in the circuit. The obtained spectrum at the output is shown in the next figure:

Figure 5.19: Power spectrum (image rejection)

Image rejection

As can be seen, the level of the output when the image signal occurs at the input is -58.69 dBm.

In the second case, only one component of the signal wasted in the circuit. In this case addition or subtraction doesn't occur. Therefore, the amplitude of the signal at the output is expected to be larger than in the first case (when "subtraction" occurs)

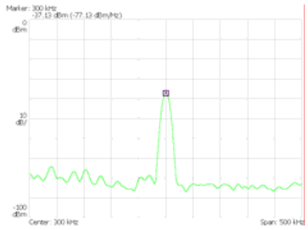


Figure 5.20: Power spectrum (only one input)

As can be seen, the level of the output when only one signal occurs at the input is -37.13 dBm.

In the third case, the in-phase and quadrature components of the wanted signal are generated by the dual arbitrary generator. A -50 dBm power level has been introduced in the circuit. The obtained spectrum at the output is shown in the next figure:

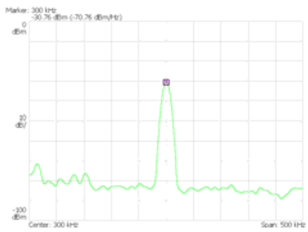


Figure 5.21: Power spectrum (wanted signal)

Wanted signal

As can be seen, the level of the output when the wanted signal occurs at the input is - 30.76 dBm.

In conclusion, the image rejection ratio of the polyphase filter can be stated by:

$$IRR) - 30.76\ \text{dBm} - (-58.69\ \text{dBm}) = 27.93\ \text{dBm} \tag{5.7}$$

Therefore, approximately 30 dB IRR are achieved with this block. Less IRR is expected when using the complete receiver chain because of the amplitude and phase errors introduced by the other blocks.

The losses of this block can be estimated as well. Taking into account the gain of the differential to single ended amplifier calculated in the previous section (32 dB), the losses of this block are -30.76 dBm (output signal) – (-50 dBm) (input signal) – 32 (gain of block 3) = 12.76 dB.

Chapter 6

Summary & Outlook

In this work the top-down design of a RF receiver for a MIMO system has been designed. The design has started with a study of the wireless channel and the MIMO channels, followed by an analysis of various receiver architectures, followed by a careful study of the receiver specification and concluded with the design, implementation and measurement of appropriate circuits.

The goal of this project was to build up a robust modular receiver to overcome the disadvantages of the previous implementation. This previous implementation is a prototype with drawbacks like asymmetric behavior. Therefore, symmetric behavior was given a high priority throughout all of the phases of the design.

In addition, in the attempt to carry out real measurements in an indoor wireless channel, it was decided at the beginning of the project to satisfy the requirements imposed by the indoor wireless channel and the measurement device.

Moreover, to enable the receiver chain to be measured with the devices available in the lab, it was wanted to build up a receiver which has a single-ended RF-interface.

The developed receiver, besides making use of a low-IF receiver architecture which allow integration of the channel filter, has a single-ended RF interface, satisfy the requirements imposed by the indoor wireless channel and the measurement device and overcome the disadvantages of the prototype because it has a symmetric behavior.

The symmetric behavior has been achieved by using only symmetric components. The robustness of the receiver has been demonstrated by means of measurements in real indoor sceneries.

Despite the good performance achieved by the receiver, a long period of optimization is required. Future works to optimize this design are:

- Improve the space occupied. The whole system can fit into a smaller space by optimizing the layout dimensions.
- Improve the isolation between the components of the system. Some components in the receiver chain can be replaced by other components which provide the circuit with more isolation. A concrete possibility already studied is to replace the power divider by a Wilkinson power divider.
- To include on board receiver antenna or local oscillator. These components are still external components in the designed receiver

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