Telecommunicatie B (EE2T21)

Telecommunicatie B bestaat uit 2 onderdelen:
- **Telecommunicatietechniek (TT)**
- **Telecommunicatienetwerken(TN)**

TT en TN hebben ieder 7 x 2 uur college.

Tentamens:
- **2 deeltentamens (ieder 2 uur, combinatie TT en TN)**
- **herkansingstentamen (3 uur, combinatie TT en TN)**

Eindcijfer = 0.6*TT + 0.4*TN

Beide deelcijfers TT, TN ≥ 5.8.

Vak halen op basis van de deeltentamens of herkansing. Een mix is niet mogelijk.

Lees info in digitale studiegids ! ! !

EE2T21 Telecommunicatie B
Dr.ir. Gerard J.M. Janssen
Dr.ir. Fernando Kuipers
April 15, 2016

Telecommunicatie B - TT (EE2T21)

Docent: Gerard Janssen Vakgroep Circuits & Systems
Room HB17.060, email: g.j.m.janssen@tudelft.nl

Boek: L.W. Couch, *"Digital and Analog Communication Systems"*,
8th edition, Prentice Hall, 2013

TC_A: Beginselen van signaaloverdracht en basisbandtransmissie
& lab courses
TC_B: Banddoorlaattransmissie: analoge en digitale modulatie technieken
& Telecommunicatienetwerken

Telecommunicatie B - TT: 7 x 2uur college
3 x 2 uur instructie

Deeltentamen 1:	woensdag 25 mei, 2016	: 13:30 – 15:30 uur
Deeltentamen 2:	woensdag 29 juni, 2016	: 13:30 – 15:30 uur
Herkansing TC_B:	maandag 25 juli, 2016	: 13:30 – 16:30 uur

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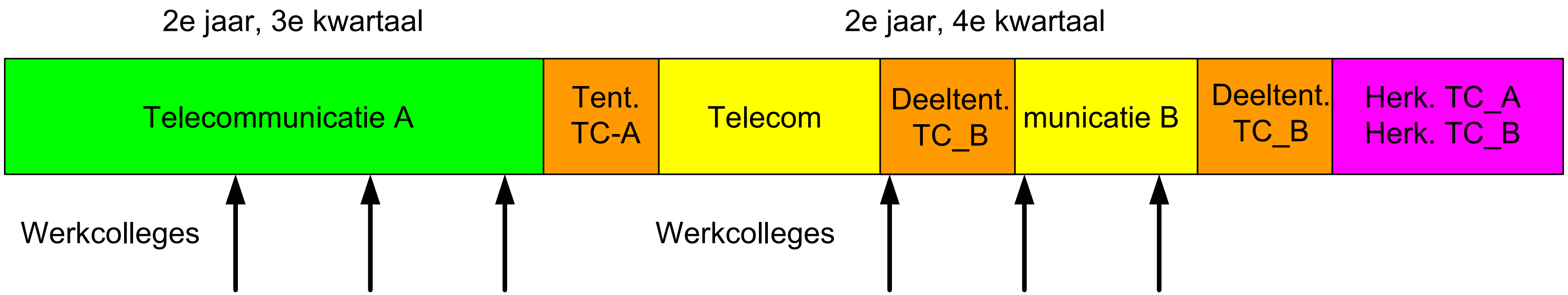
Telecommunicatie A en B



Telecommunicatie A

De theoretische basis van communicatiesystemen: signaalbeschrijving, signaalpropagatie, ruis en systeemruisberekeningen, basisbandtechnieken, modulatietheorie → analoge pulsmodulatie, analoog/digitaalomzetting.

& lab courses

Telecommunicatie B - TT

Banddoorlaatsystemen: modulatietechnieken voor analoge en digitale signalen. Maten voor de kwaliteit van detectie van signalen met ruis: berekening van de signaal-ruis verhouding en bitfoutenkans.

Collegerooster (1)

Collegestof Telecommunicatie B EE2T21 bij de 8e editie van Couch **4e kwartaal 2015-2016**

Col./Instr	Datum	Hfdst	Onderwerp	Blz.
Col. 1	18-4		**Banddoorlaatsignalen en systemen**	
		4.1 - 4.7	Signal descriptions:	
			bandpass signals and systems, modulation	259 - 276
		4.16	Transmitters and receivers	312 - 320
Col. 2	25-4		**Lineaire analoge modulatievormen**	
		5.1 - 5.5	AM, DSB-SC and SSB modulation	335 - 350
		4.13	Detector circuits (envelope and product detector)	296 - 299
Col. 3	2-5		**Niet-lineaire analoge modulatievormen**	
		7.8	Output SNR for analog modulations:	
			AM, DSB-SC and SSB	552 - 558
		5.6	Phase and frequency modulation	318 - 333
Col. 4	9-5		**Digitale modulatievormen (1)**	
		7.8	Output SNR for analog modulations: PM, FM	558 - 565
		7.9	Comparison of analog systems	570 - 572
		5.9	Bandpass communication system, digital communication,	
			binary modulation techniques: OOK, BPSK, DPSK, FSK,	375 - 388
Col. 5	10-5		**Digitale modulatievormen (2)**	
		5.10 - 5.11	Multilevel modulation: QPSK, π/4-QPSK QAM, MSK	388 - 407
		5.12	Orthogonal Frequency Division Multiplexing (OFDM)	407 - 410
		5.13	Spread spectrum systems: Direct Sequence (DS-SS)	410 - 417
Werkcollege 1	17-5			

Blackboardpagina EE2T21, tab “Course information”

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Collegerooster (2)

Collegestof Telecommunicatie B EE2T21 bij de 8e editie van Couch **4e kwartaal 2015-2016**

Col./Instr	Datum	Hfdst	Onderwerp	Blz.
Deeltentamen 1	25-5			
Col. 6	30-5		**Bitfoutenkans in digitale transmissiesystemen (1)**	
		6.8	Matched filters	486 - 494
		7.1	Bit Error Probability	514 - 521
		7.2	BER in baseband systems	521 - 526
Werkcollege 2	31-5			
Col. 7	6-6		**Bitfoutenkans in digitale transmissiesystemen (2)**	
		7.3	Coherent demodulation: OOK, BPSK, FSK	526 - 533
		7.4	Non-coherent demodulation: DPSK	533 - 540
		7.5	QPSK, π/4-QPSK, MSK	541 - 543
		7.6	Comparison of digital systems	543 - 547
Werkcollege 3	13-6			
Deeltentamen 2	29-6			

Tentaminering

Deeltentamen 1:	woensdag 25 mei 2016, 13:30 – 15:30 uur	Stof: colleges 1-3 plus college 4: par. 7.8 en 7.9.
Deeltentamen 2:	woensdag 29 juni 2016, 13:30 – 15:30 uur	Stof: digitale modulatievormen en bitfoutenkans: coll. 4 (par. 5.9) en coll. 5-7.
Herkansing:	maandag 25 juli 2016, 13:30 – 16:30 uur	

Bij de deeltentamens voor Telecommunicatietechniek mogen het boek van Couch en een niet-programmeerbare rekenmachine gebruikt worden. U dient zich voor de beide deeltentamens aan te melden via OSIRIS.

Blackboardpagina EE2T21, tab “Course information”

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Colleges en Instructies
Telecommunicatie B

Colleges:

Maandag	18-4, 25-4, 2-5, 9-5	5e+6e uur, EWI-CZ Chip
	30-5, 6-6	
Dinsdag	10-5	7e+8e uur, EWI-CZ Pi

Instructies:

Dinsdag	17-5	5e+6e uur, EWI-CZ Boole
Dinsdag	31-5	7e+8e uur, EWI-CZ Pi
Maandag	13-6	5e+6e uur, EWI-CZ Chip

Huishoudelijk

- Mededelingen komen op de Blackboard pagina van het vak Telecommunicatie B (EE2T21)

- Overzicht van de collegestof-per-week op Blackboard

- Wekelijks verschijnt er een MapleTA huiswerkopgave op Blackboard behorende bij de stof van die week.

- College responsiegroep

"Bonus" verdienen met de Huiswerkopdrachten

Maken van de MapleTA huiswerkopdrachten levert een bonus op voor het Telecommunicatietechniek (TT) onderdeel.

Voor Telecommunicatienetwerken (TN) is ook een bonus te verdienen.

$$\text{Cijfer TT} = X + 2*HW*(10-X)/100$$

met X = {gemiddelde deeltentamens (DTT1 + DTT2)/2, herkansing}.

Lees info in digitale studiegids ! ! !

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Huiswerkopdrachten met MapleTA (1)

De MapleTA opgaven staan onder "Assignments".

- Je kunt de opgaven printen en vervolgens eerst maken voordat je de antwoorden in MapleTA invult
- Als de opdracht uit meerdere vragen bestaat, vul je eerst alle antwoorden in voordat je de "Grade" knop gebruikt.
- "Grade" is:
  * antwoorden inleveren,
  * je krijgt een cijfer
  * einde opdracht.
- Je kunt de opdracht maar één keer maken!!!
- Als je met Quit/Save de opdracht verlaat kun je later terugkomen.

Huiswerkopdrachten met MapleTA (2)

- De vragen staan tenminste 1 week open:
  * vanaf kort na het college
  * tot 's ochtends 3:00 uur op de dag van het eerstvolgende college
- Geef de antwoorden op twee decimalen nauwkeurig.

- **LET OP!!! Het decimaalteken in MapleTA is een "PUNT".**

- Er wordt nagekeken met een absolute antwoordnauwkeurigheid van +/- 0.2 dB, of een relatieve nauwkeurigheid van +/- 2%

Telecommunicatie B - TT (EE2T21)

Lecture 8 overview:

General

Bandpass signals and systems

*** modulation: why and how?**

*** mathematical description of bandpass signals and systems**

*** bandpass sample theorem**

Basic transmitter and receiver principles

*** transmitter: AM/PM, In-phase/Quadrature phase (IQ)**

*** receivers: tuned radio frequency receiver, superheterodyne receiver, homodyne receiver**

EE2T21 Telecommunicatie B - TT

Dr.ir. Gerard J.M. Janssen

April 15, 2016

Aim of Telecommunicatie B - TT

In this course, we study bandpass communication systems:
- description of bandpass signals and systems
- transmitter and receiver concepts
- analogue and digital modulation techniques,

  and their performance in the presence of Additive White Gaussian Noise (AWGN).

We define quality measures for the received signal after detection:

- SNR $\Rightarrow$ analogue modulation
- BER $\Rightarrow$ digital modulation

Modulation (1)

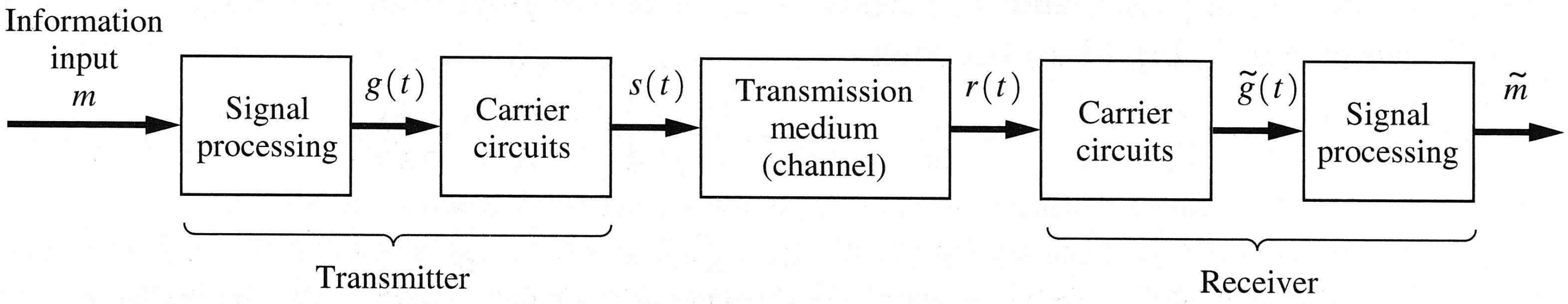


Figure 4–1 Communication system.

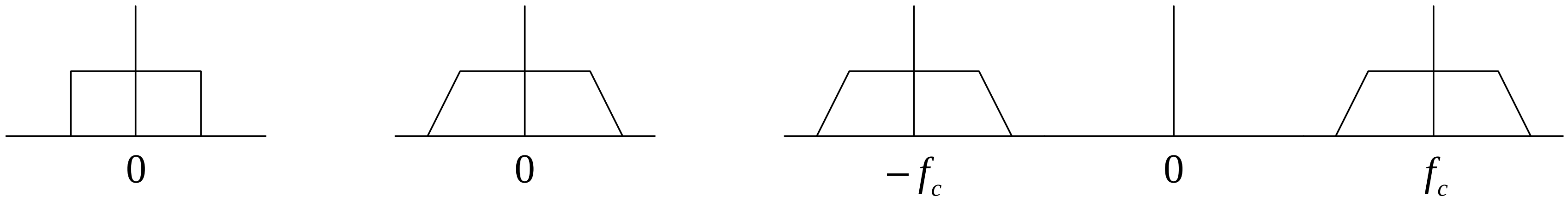
What is modulation?

Modulation is the method to adapt one or more carrier signal parameters according to an information signal, in order to make it suitable for transmission over the available channel.

Modulation (2)

baseband signal $\leftrightarrow$ (de-)modulation $\leftrightarrow$ bandpass signal

$m(t)$ $\quad g(t) = f(m(t))$ $\quad s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\}$



Baseband signals: $|f| \leq B$

Bandpass signals: $f_{\mathrm{min}} \leq |f| \leq f_{\mathrm{max}}$

Why modulation?

1. Not enough space in baseband: distortion, interference
2. Baseband signaling is not suitable for every medium?
3. Signals can be stacked in the frequency domain:
   FDM = Frequency Division Multiplexing
4. Signals can be easily separated in the frequency domain

Bandpass Signals (1)

Every physical signal is real!

Mathematical description of bandpass signals:

$$s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\}$$
$$= R(t)\cos[\omega_c t + \theta(t)] = x(t)\cos\omega_c t - y(t)\sin\omega_c t$$

$$x(t) = R(t)\cos\theta(t) \qquad R(t) = |g(t)| = \sqrt{x^2(t) + y^2(t)}$$

$$y(t) = R(t)\sin\theta(t) \qquad \theta(t) = \arctan\frac{y(t)}{x(t)}$$

Complex envelope or complex equivalent baseband signal:

$$g(t) = f(m(t)) = x(t) + jy(t) \quad = \quad R(t)e^{j\theta(t)}$$

Modulation function **In-phase and Quadrature-phase component** **Amplitude: AM** **Phase: PM, FM**

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Bandpass Signals (2)

The amplitude:

$$R(t) = |g(t)| = \sqrt{x^2(t) + y^2(t)} \geq 0$$: real

The phase:

$$\theta(t) = \arctan \frac{y(t)}{x(t)}$$: real

$g(t),\ x(t),\ y(t),\ R(t)$ and $\theta(t) \rightarrow$ baseband signals

$s(t) = \mathrm{Re}\{g(t) e^{j\omega_c t}\}$ $\rightarrow$ transformation from baseband to a bandpass signal

$g(t) = f(m(t))$ $\rightarrow$ modulation function determined by the selected modulation technique

Modulation techniques (1)

TABLE 4–1 COMPLEX ENVELOPE FUNCTIONS FOR VARIOUS TYPES OF MODULATION[a]

Type of Modulation	Mapping Functions $g(m)$	Corresponding Quadrature Modulation $x(t)$	$y(t)$
AM	$A_c[1+m(t)]$	$A_c[1+m(t)]$	0
DSB-SC	$A_c m(t)$	$A_c m(t)$	0
PM	$A_c e^{jD_p m(t)}$	$A_c \cos[D_p m(t)]$	$A_c \sin[D_p m(t)]$
FM	$A_c e^{jD_f \int_{-\infty}^{t} m(\sigma)\, d\sigma}$	$A_c \cos\left[D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma\right]$	$A_c \sin\left[D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma\right]$
SSB-AM-SC[b]	$A_c[m(t) \pm j\hat{m}(t)]$	$A_c m(t)$	$\pm A_c \hat{m}(t)$
SSB-PM[b]	$A_c e^{jD_p[m(t) \pm j\hat{m}(t)]}$	$A_c e^{\mp D_p \hat{m}(t)} \cos[D_p m(t)]$	$A_c e^{\mp D_p \hat{m}(t)} \sin[D_p m(t)]$
SSB-FM[b]	$A_c e^{jD_f \int_{-\infty}^{t}[m(\sigma) \pm j\hat{m}(\sigma)]d\sigma}$	$A_c e^{\mp D_f \int_{-\infty}^{t} \hat{m}(\sigma)d\sigma} \cos\left[D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma\right]$	$A_c e^{\mp D_f \int_{-\infty}^{t} \hat{m}(\sigma)d\sigma} \sin\left[D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma\right]$
SSB-EV[b]	$A_c e^{\{\ln[1+m(t)] \pm j\hat{\ln}\|1+m(t)\|\}}$	$A_c[1+m(t)] \cos\{\hat{\ln}[1+m(t)]\}$	$\pm A_c[1+m(t)] \sin\{\hat{\ln}[1+m(t)]\}$
SSB-SQ[b]	$A_c e^{(1/2)\{\ln[1+m(t)] \pm j\hat{\ln}\|1+(t)\|\}}$	$A_c\sqrt{1+m(t)} \cos\{\frac{1}{2}\hat{\ln}[1+m(t)]\}$	$\pm A_c\sqrt{1+m(t)} \sin\{\frac{1}{2}\hat{\ln}[1+m(t)]\}$
QM	$A_c[m_1(t)+jm_2(t)]$	$A_c m_1(t)$	$A_c m_2(t)$

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Modulation techniques (2)

TABLE 4–1 COMPLEX ENVELOPE FUNCTIONS FOR VARIOUS TYPES OF MODULATION (*cont.*)

Type of Modulation	Corresponding Amplitude and Phase Modulation: $R(t)$	$\theta(t)$	Linearity	Remarks
AM	$A_c\|1 + m(t)\|$	$\begin{cases} 0, & m(t) > -1 \\ 180°, & m(t) < -1 \end{cases}$	L[c]	$m(t) > -1$ required for envelope detection
DSB-SC	$A_c\|m(t)\|$	$\begin{cases} 0, & m(t) > 0 \\ 180°, & m(t) < 0 \end{cases}$	L	Coherent detection required
PM	A_c	$D_p m(t)$	NL	D_p is the phase deviation constant (rad/volt)
FM	A_c	$D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma$	NL	D_f is the frequency deviation constant (rad/volt-sec)
SSB-AM-SC[b]	$A_c\sqrt{[m(t)]^2 + [\hat{m}(t)]^2}$	$\tan^{-1}[\pm\hat{m}(t)/m(t)]$	L	Coherent detection required
SSB-PM[b]	$A_c e^{\pm D_p \hat{m}(t)}$	$D_p m(t)$	NL	
SSB-FM[b]	$A_c e^{\pm D_f \int_{-\infty}^{t} \hat{m}(\sigma) d\sigma}$	$D_f \int_{-\infty}^{t} m(\sigma)\, d\sigma$	NL	
SSB-EV[b]	$A_c\|1 + m(t)\|$	$\pm \hat{\ln}[1 + m(t)]$	NL	$m(t) > -1$ is required so that $\ln(\cdot)$ will have a real value
SSB-SQ[b]	$A_c\sqrt{1 + m(t)}$	$\pm\frac{1}{2} \hat{\ln}[1 + m(t)]$	NL	$m(t) > -1$ is required so that $\ln(\cdot)$ will have a real value
QM	$A_c\sqrt{m_1^2(t) + m_2^2(t)}$	$\tan^{-1}[m_2(t)/m_1(t)]$	L	Used in NTSC color television; requires coherent detection

[a] $A_c > 0$ is a constant that sets the power level of the signal as evaluated by the use of Eq. (4–17); L, linear; NL, nonlinear; $[\hat{\ }]$ is the Hilbert transform (i.e., the $-90°$ phase-shifted version) of $[\cdot]$. (See Sec. 5–5 and Sec. A–7, Appendix A.)

[b] Use upper signs for upper sideband signals and lower signs for lower sideband signals.

[c] In the strict sense, AM signals are not linear, because the carrier term does not satisfy the linearity (superposition) condition.

Bandpass signal characteristics (1)

Amplitude spectrum:

$$
\begin{aligned}
s(t) &= R(t)\cos[\omega_c t + \theta(t)] = x(t)\cos\omega_c t - y(t)\sin\omega_c t \\
&= \mathrm{Re}\{g(t)e^{j\omega_c t}\} = \tfrac{1}{2}[g(t)e^{j\omega_c t} + g^*(t)e^{-j\omega_c t}]
\end{aligned}
$$

$$
\begin{aligned}
S(f) = \mathfrak{F}\{s(t)\} &= \mathfrak{F}\{\tfrac{1}{2}g(t)e^{j\omega_c t} + \tfrac{1}{2}g^*(t)e^{-j\omega_c t}\} \\
&= \tfrac{1}{2}[G(f - f_c) + G^*(-f - f_c)]
\end{aligned}
$$

where we used:

$$
\begin{aligned}
G(f) &= \mathfrak{F}\{g(t)\} \\
G^*(-f) &= \mathfrak{F}\{g^*(t)\}
\end{aligned}
$$

Bandpass signal characteristics (2)

Power Spectral Density (PSD):

$$P_s(f) = \tfrac{1}{4}P_g(f - f_c) + \tfrac{1}{4}P_g(-f - f_c) = |S(f)|^2 \quad \text{[W/Hz]}$$

where we used: $P_g(f) = PSD\{g(t)\} = |G(f)|^2$

Average signal power (normalized to a 1Ω load):

$$P_s = \langle s^2(t) \rangle = \tfrac{1}{2}\langle |g(t)|^2 \rangle = \int_{-\infty}^{\infty} P_s(f)\,df$$

due to sinewave carrier

Peak Envelope Power (PEP): the average power when $|g(t)|$ is maximum

Normalized PEP: $P_{s\_PEP} = \tfrac{1}{2}[\max\{|g(t)|\}]^2$

Example: Amplitude Modulation (1)

Complex envelope: $g(t) = A_c[1 + m(t)]$ with $m(t)$ real.

Note that the carrier amplitude A_c is part of $g(t)$.

AM-signal: $s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\} = A_c[1 + m(t)]\cos\omega_c t$

Spectrum $g(t)$: $G(f) = \mathfrak{F}\{g(t)\} = A_c\delta(f) + A_c M(f) \;\rightarrow\; M(f) = \mathfrak{F}\{m(t)\}$

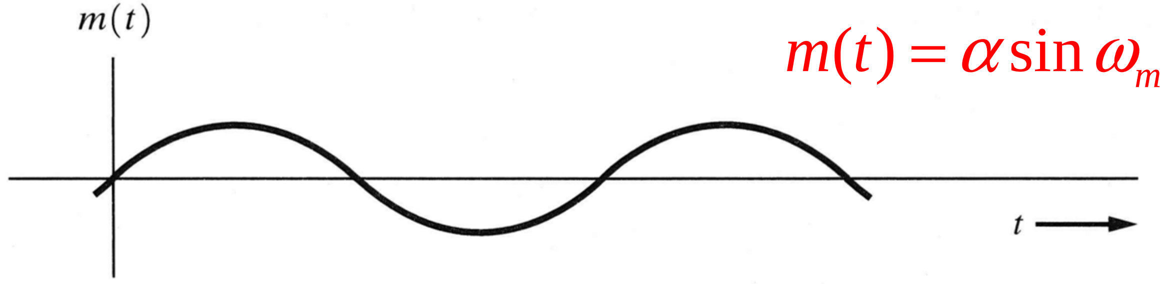
Spectrum $s(t)$: $S(f) = \mathfrak{F}\{A_c[1 + m(t)] \cdot \tfrac{1}{2}[e^{j\omega_c t} + e^{-j\omega_c t}]\}$

$$= \tfrac{1}{2}A_c[\delta(f - f_c) + M(f - f_c) + \delta(f + f_c) + M(f + f_c)]$$

where we used that: $M^*(-f) = M(f)$ since $m(t)$ is real.

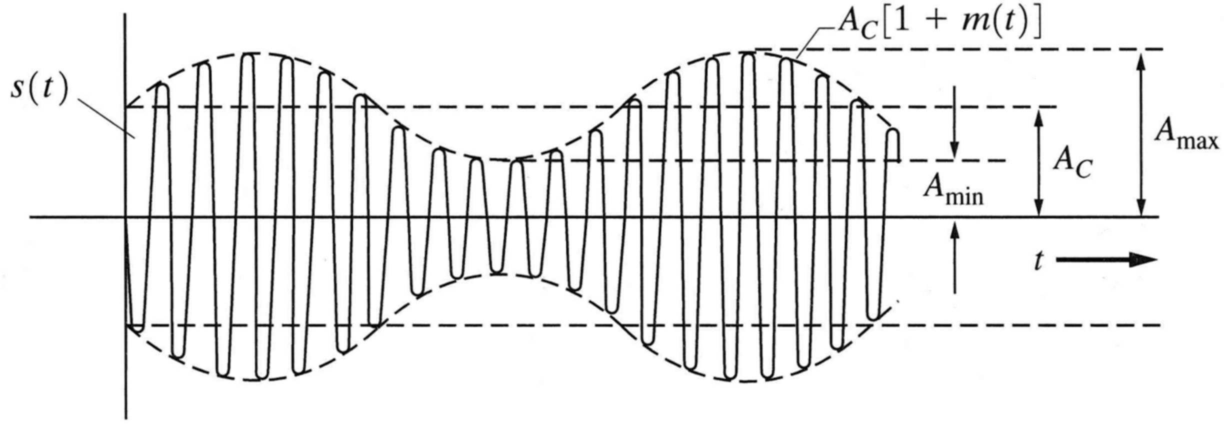
Example: Amplitude Modulation (2)

$$m(t) = \alpha \sin \omega_m t$$



(a) Sinusoidal Modulating Wave

$$g(t) = A_c[1 + m(t)] = A_c[1 + \alpha \sin \omega_m t]$$



(b) Resulting AM Signal

$$s(t) = A_c[1 + \alpha \sin \omega_m t] \cos \omega_c t$$

Figure 5–1 AM signal waveform.

Example: Amplitude Modulation (3)

Average normalized power:

$$P_{s\_AM} = \langle s^2(t) \rangle = \tfrac{1}{2} A_c^2 \langle [1+m(t)]^2 \rangle$$

$$= \tfrac{1}{2} A_c^2 \langle [1+2m(t)+m^2(t)] \rangle$$

$$= \tfrac{1}{2} A_c^2 + \tfrac{1}{2} A_c^2 \langle m^2(t) \rangle$$

$$= P_{carrier} + P_{sidebands}$$

Under the assumption that $m(t)$ does not contain a DC component:

$$\Rightarrow \quad \langle m(t) \rangle \quad = 0$$

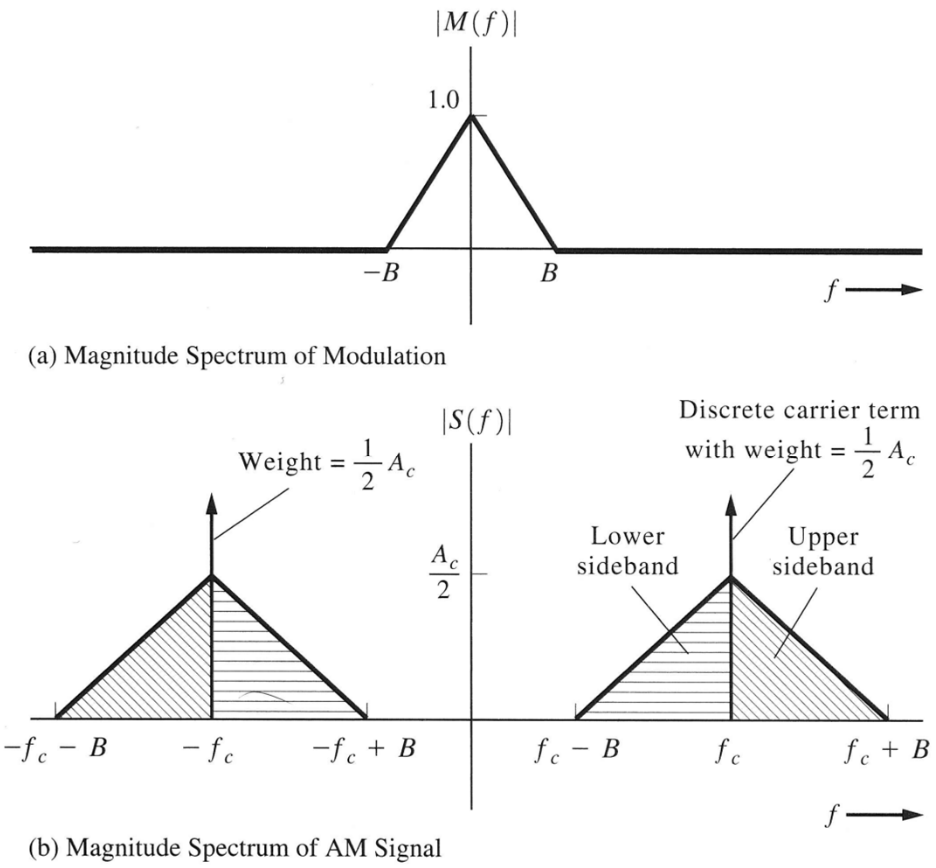
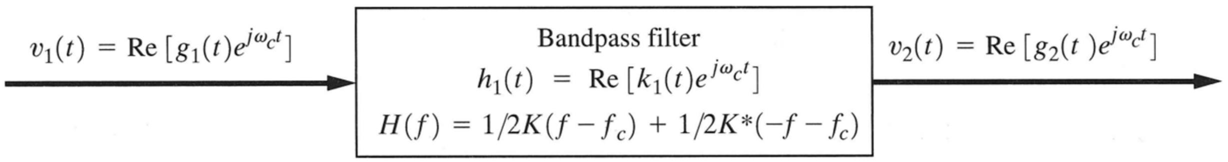


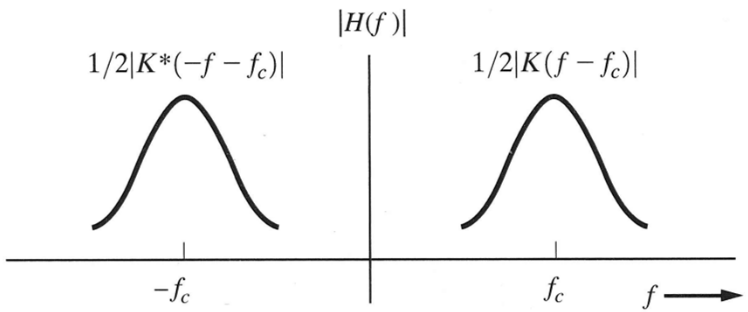
Figure 4–2 Spectrum of AM signal.

Bandpass filtering (1)

Like bandpass signals, bandpass systems can also be modeled using the complex equivalent baseband / complex envelope representation.



(a) Bandpass Filter

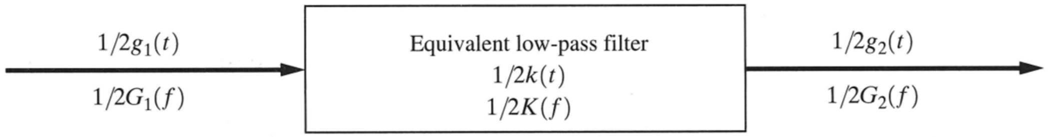


(b) Typical Bandpass Filter Frequency Response

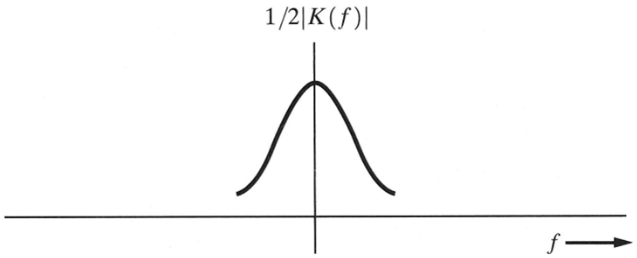
$$h(t) = \mathrm{Re}\{k(t)e^{j\omega_c t}\} = \tfrac{1}{2}[k(t)e^{j\omega_c t} + k^*(t)e^{-j\omega_c t}] \quad \leftrightarrow \quad \tfrac{1}{2}k(t)$$

$$H(f) = \tfrac{1}{2}K(f-f_c) + \tfrac{1}{2}K^*(-f-f_c) \quad \leftrightarrow \quad \tfrac{1}{2}K(f)$$

Bandpass filtering (2)



(c) Equivalent (Complex Impulse Response) Low-pass Filter



(d) Typical Equivalent Low-pass Filter Frequency Response

Figure 4–3 Bandpass filtering.

$$\tfrac{1}{2}g_2(t) = \tfrac{1}{2}g_1(t) * \tfrac{1}{2}k(t)$$

$$\tfrac{1}{2}G_2(f) = \tfrac{1}{2}G_1(f) \cdot \tfrac{1}{2}K(f)$$

Why?

Usually it is much simpler to do calculations in the baseband domain than in the bandpass domain. This is what actually happens in digital signal processing and simulation software.

Distortion in bandpass filtering

AM / PM in the input signal of a bandpass filter can cause PM / AM in the output signal:

$$R_1(t) = |g_1(t)| \quad \rightarrow \quad \theta_2(t) = ang\{g_2(\mathrm{t})\} \quad : \text{ AM-PM conversion}$$

$$\theta_1(t) \quad \rightarrow \quad R_2(t) \quad : \text{ PM-AM conversion}$$

$R_2(t)$ and $\theta_2(t)$ are non-linear filtered versions of $x_1(t)$ and $y_1(t)$ due to $R(t) = \sqrt{x^2(t) + y^2(t)}$ and $\theta(t) = \arctan\frac{y(t)}{x(t)}$, whereas $g_2(t)$ is a linear filtered version of $g_1(t)$ using $h(t) \leftrightarrow H(f)$.

Determining distortion for bandpass systems (i.e. with modulation) requires complex calculations.

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Distortion-less bandpass transmission (1)

We wish distortion-less transmission of bandpass signals.
What are the requirements for a bandpass filter $H(f) = |H(f)|e^{j\theta(f)}$, representing a channel (guided or wireless), for distortion-less signal transmission?

1. the amplitude is constant over the passband:

$$|H(f)| = A \geq 0 \qquad \text{(4-27a)}$$

2. the derivative of the phase $d\theta / df$ is constant over the passband:

$$-\frac{1}{2\pi}\frac{d\theta(f)}{df} = -\frac{d\theta(\omega)}{d\omega} = T_g \quad \Rightarrow \quad \text{constant group delay} \qquad \text{(4-27b)}$$

These requirements are weaker than for distortion-less transmission in baseband as discussed in §2.6, since:

$$\theta(f) = -2\pi f T_g + \theta_0$$

Bandpass

$$\theta(f) = -2\pi f T_d$$

Baseband

TU Delft

Distortion-less bandpass transmission (4)

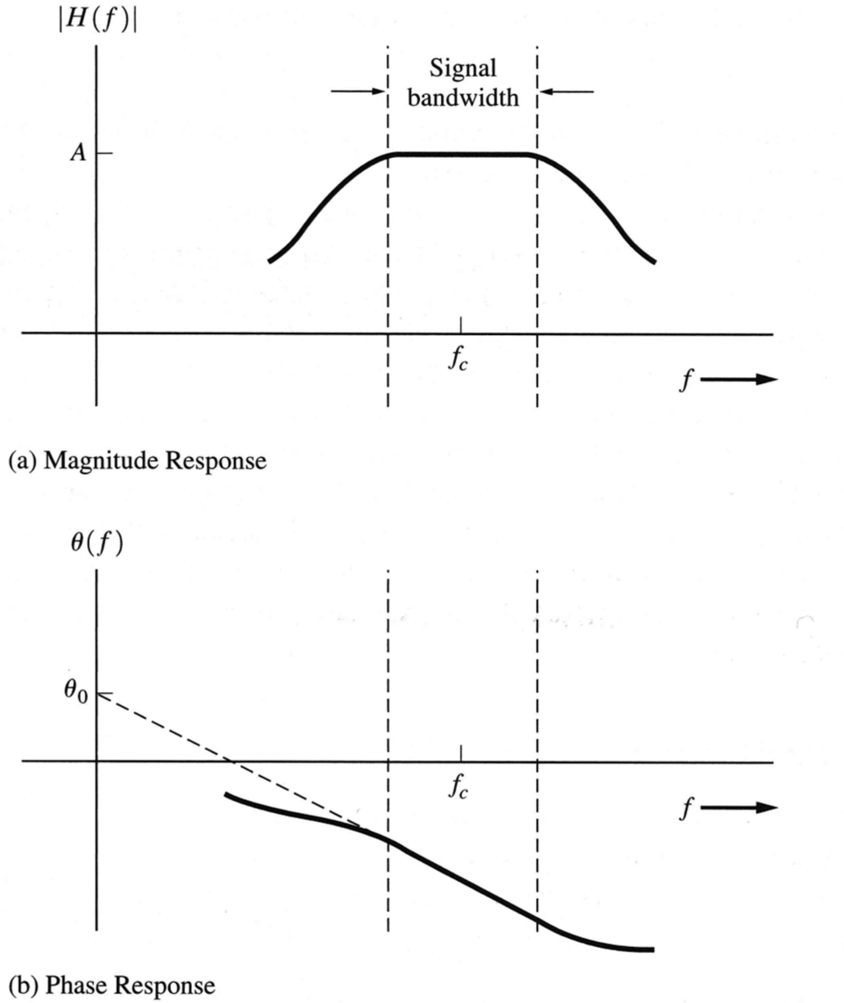


Figure 4–4 Transfer characteristics of a distortionless bandpass channel.

Distortion-less bandpass transmission (2)

Proof that 4-27a and 4-27b are sufficient for distortion-less transmission of bandpass signals. Choose:

$$H(f) = A\exp\{j(-2\pi fT_g + \theta_0\}$$

$$= \underbrace{A\exp\{j\theta_0\}}_{\text{complex amplitude}} \cdot \underbrace{\exp\{-j(2\pi fT_g\}}_{\text{delay of the information signal}}$$

Now the input signal: $v_1(t) = x(t)\cos\omega_c t - y(t)\sin\omega_c t$

results in:

$$v_2(t) = Ax(t-T_g)\cos(\omega_c(t-T_g)+\theta_0) - Ay(t-T_g)\sin(\omega_c(t-T_g)+\theta_0)$$

$$= Ax(t-T_g)\cos(\omega_c t+\theta(f_c)) - Ay(t-T_g)\sin(\omega_c t+\theta(f_c))$$

Distortion-less bandpass transmission (3)

For distortion-less transmission in baseband, the signal's phase should be a linear decreasing function of frequency (2-150b):

$$\theta(f_c) \triangleq -\omega_c T_g + \theta_0 \overset{requirement}{=} -2\pi f_c T_d \qquad T_d = \text{phase delay time}$$

Now we can write:

$$v_2(t) = Ax(t-T_g)\cos\omega_c(t-T_d) - Ay(t-T_g)\sin\omega_c(t-T_d)$$

The group delay T_g $\neq$ phase delay T_d for bandpass signals (when $\theta_0 \neq 0$). However, this is not a problem, because the information is not in the carrier signal but in its modulation.

Bandpass sampling theorem (1)

According to Nyquist we find for the sample frequency:

$f_s \geq 2B$ $\quad$ B is the highest signal frequency

For a modulated bandpass signal this would result in:

$f_s \geq 2\left(f_c + \frac{B_T}{2} \right)$ where $B_T = f_{\max} - f_{\min}$ is the transmission bandwidth.

Fortunately, there is no need to sample bandpass signals with twice the highest signal frequency in order to capture its information. Note that the information is contained in the transmission bandwidth B_T and not in the carrier frequency.

Therefore, the minimum required sample frequency for bandpass signal is:

$$f_s \geq 2B_T$$

Bandpass sampling theorem (2)

Proof: let $v(t) = x(t)\cos\omega_c t - y(t)\sin\omega_c t$ with $f_c = (f_{\min} + f_{\max})/2$ and $x(t)$, $y(t)$ baseband signals with absolute bandwidth $B = B_T/2$.

We know baseband sampling: $x(t)$ and $y(t)$ are sampled with $f_b \geq 2B = B_T$.
However, here $x(t)$ as well as $y(t)$ need to be sampled so: $f_s \geq 2B_T$

Now we can write:

$$v(t) = \sum_{n=-\infty}^{\infty} \left[x(nT_b)\cos\omega_c t - y(nT_b)\sin\omega_c t\right] \underbrace{\frac{\sin \pi f_b (t - nT_b)}{\pi f_b (t - nT_b)}}_{\text{reconstruction filter}}$$

The frequency: $f_s = 2 f_b = 2B_T$ is called the bandpass sampling frequency.

Bandpass sampling theorem (3)

Now let us choose $t_{x,n},\ t_{y,n}$ such that:

1. $t_{x,n} = nT_b \quad \Rightarrow \quad \cos\omega_c t_{x,n} = 1,\ \sin\omega_c t_{x,n} = 0$

$$\Rightarrow \quad v(t_{x,n}) = x(nT_b)$$

2. $t_{y,n} = nT_b + \dfrac{1}{4f_c} \quad \Rightarrow \quad \cos\omega_c t_{y,n} = 0,\ \sin\omega_c t_{y,n} = 1$

$$\Rightarrow \quad v(t_{y,n}) \cong y(nT_b)$$

So we take two duo-samples with a time separation $1/4f_c$ and a period T_b.

This is correct when $f_c >> B_T \quad \Rightarrow \quad x(t),\ y(t)$ hardly change over $\Delta t = 1/4f_c$.

Or quadrature-demodulation can be applied with $\cos\omega_c t$ and $\sin\omega_c t$ followed by sampling the output signals $x(t)$ and $y(t)$ at nT_b.

I-Q-detector

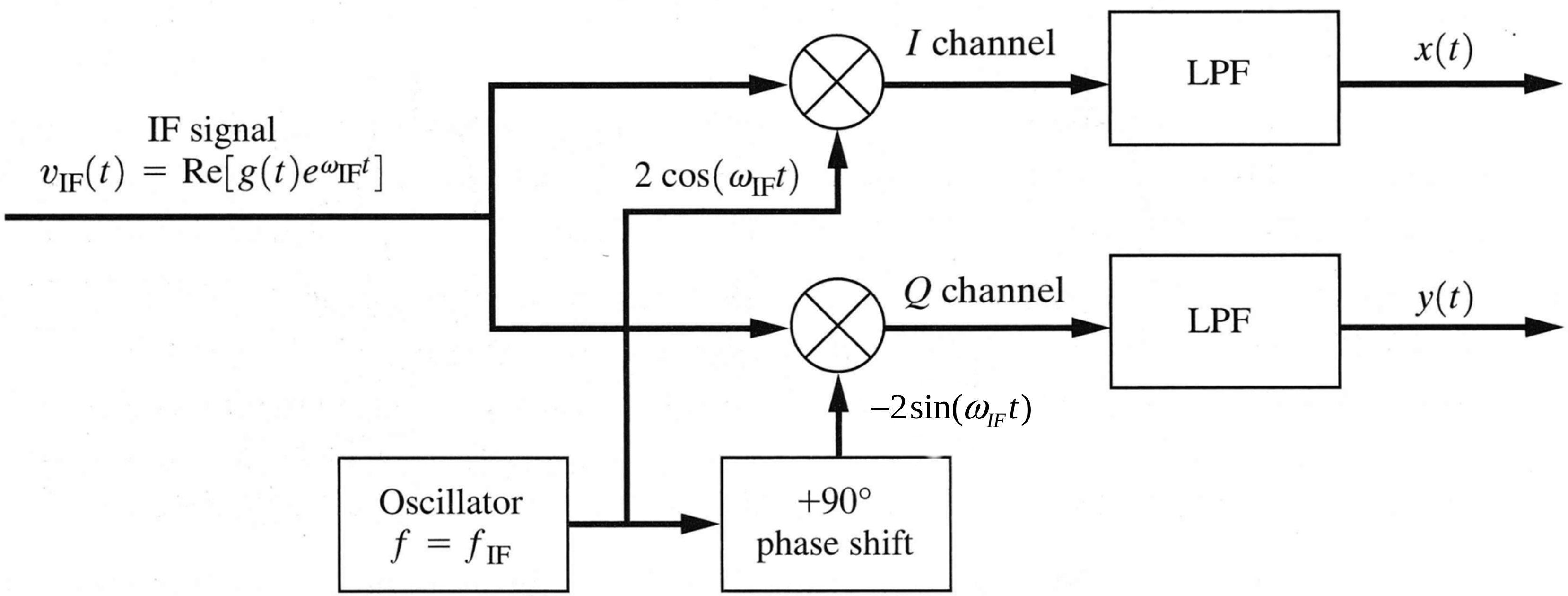


Figure 4–31 IQ (in-phase and quadrature-phase) detector.

Bandpass Dimensionality theorem

A bandpass signal with transmission bandwidth $B_T << f_c$ is fully determined over a period T_0 by:

$$N = 2B_T T_0$$

independent sample values and N is the dimension of the signal.

Signals and noise

Bandpass signals and noise both can be described using the complex equivalent baseband notation:

Transmitted signal:

$$s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\}$$
$$= |g(t)|\cos[\omega_c t + \theta_g(t)] = x(t)\cos\omega_c t - y(t)\sin\omega_c t$$

Received signal:

$$r(t) = s(t) * h(t) + n(t)$$

with: $$n(t) = \underbrace{n_x}_{\text{In-phase noise}}(t)\cos\omega_c t - \underbrace{n_y}_{\text{Quadrature-phase noise}}(t)\sin\omega_c t$$

Transmission System

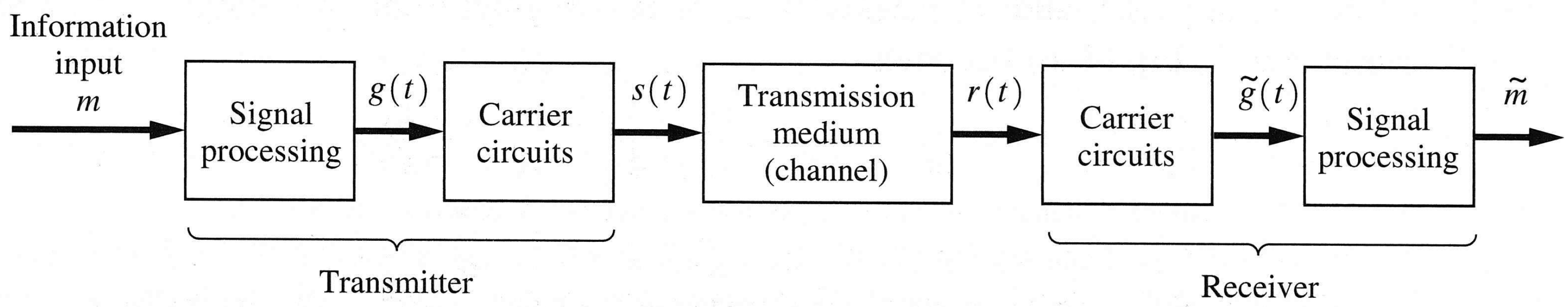


Figure 4–1 Communication system.

Basic components of a transmission system:

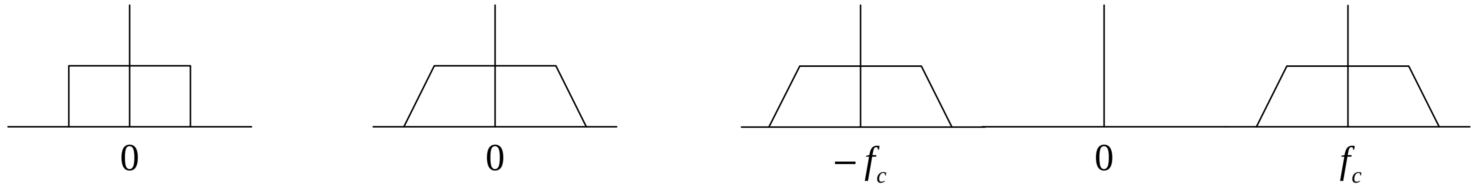
- Source
- Transmitter
- Channel
- Receiver
- Destination

Transmitters (1)

The transmitter transforms the information signal $m(t)$ into a modulated carrier signal $s(t)$ at frequency f_c which is suitable for the available transmission channel.

baseband signal $\rightarrow$ modulation $\rightarrow$ bandpass signal

$m(t)$ $\quad g(t) = f\{m(t)\}$ $\quad s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\}$



AM PM

$$s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\} = \begin{cases} R(t)\cos\left(\omega_c t + \theta(t)\right) & \text{AM-PM modulation} \\ x(t)\cos\omega_c t - y(t)\sin\omega_c t & \text{I\&Q modulation} \end{cases}$$

In-phase Quadrature-phase

We distinguish two generic transmitter concepts, based on:

1. AM-PM modulation;
2. I&Q modulation.

TU Delft

Transmitters (2)

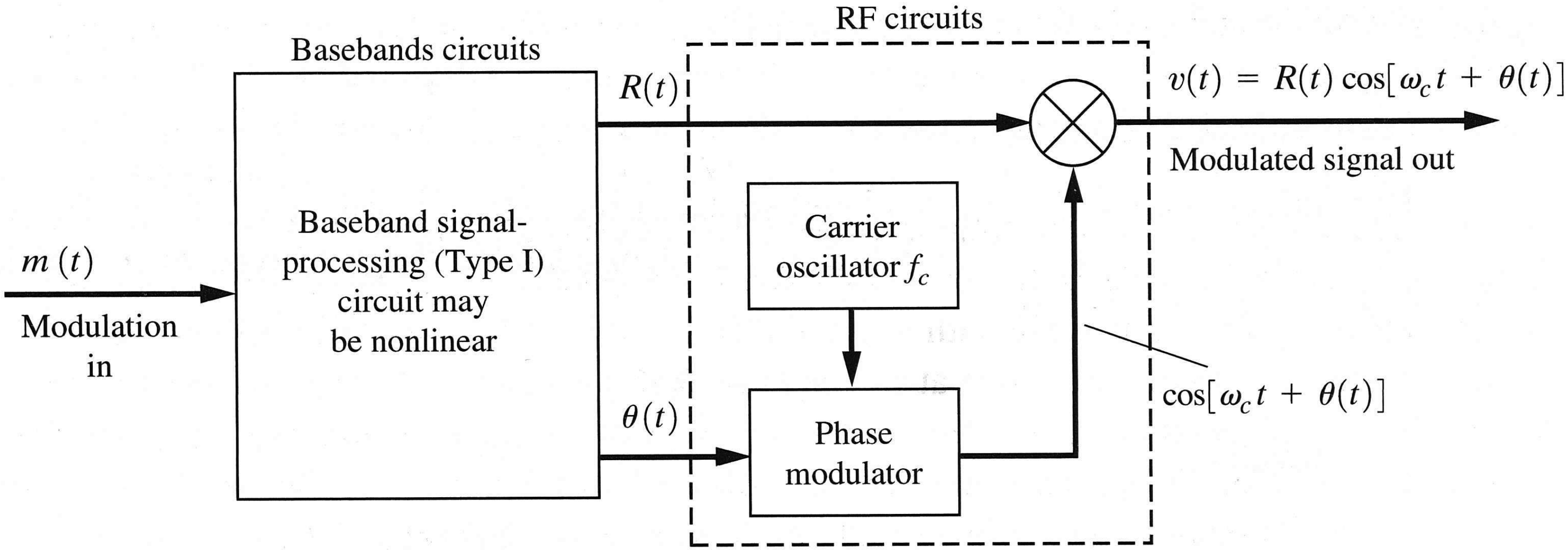


Figure 4–27 Generalized transmitter using the AM–PM generation technique.

Where:

$$R(t) = |g(t)| = \sqrt{x^2(t) + y^2(t)}$$

$$\theta(t) = \text{angle}\{g(t)\} = \arctan\frac{y(t)}{x(t)}$$

Transmitters (3)

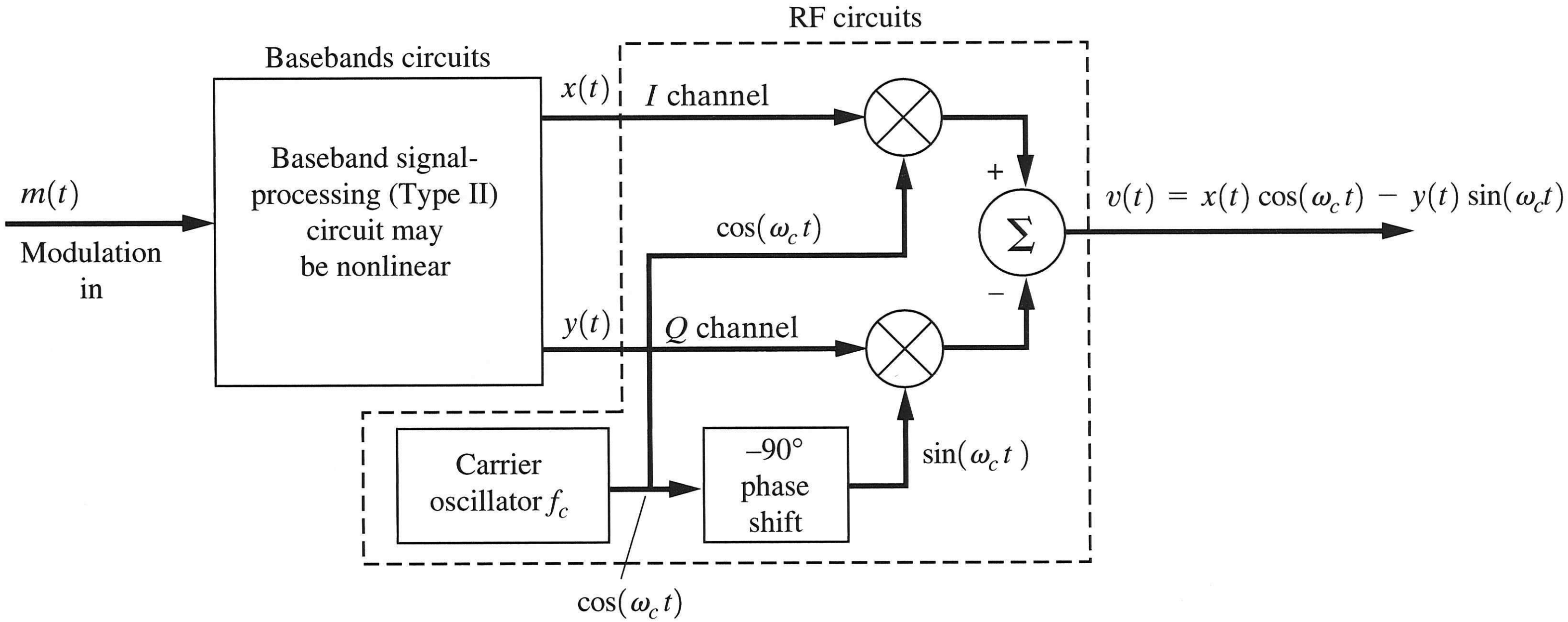


Figure 4–28 Generalized transmitter using the quadrature generation technique.

Where: $x(t) = R(t)\cos\theta(t)$

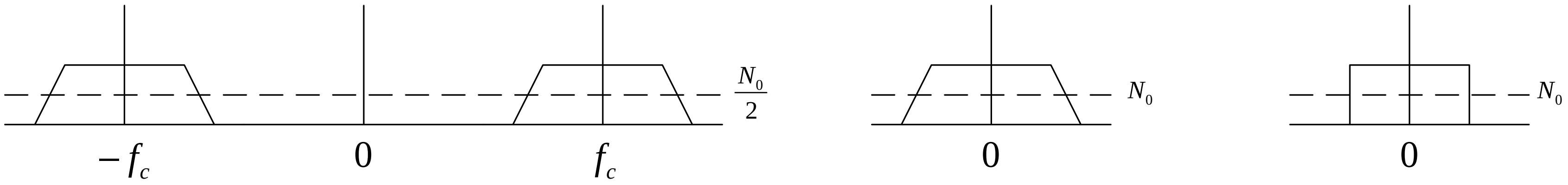
$$y(t) = R(t)\sin\theta(t)$$

Receivers

The aim of the receiver is to retrieve the transmitted information signal from the received bandpass signal (+ noise and interference) as accurately as possible at the destination.

bandpass signal $\rightarrow$ demodulation $\rightarrow$ baseband signal

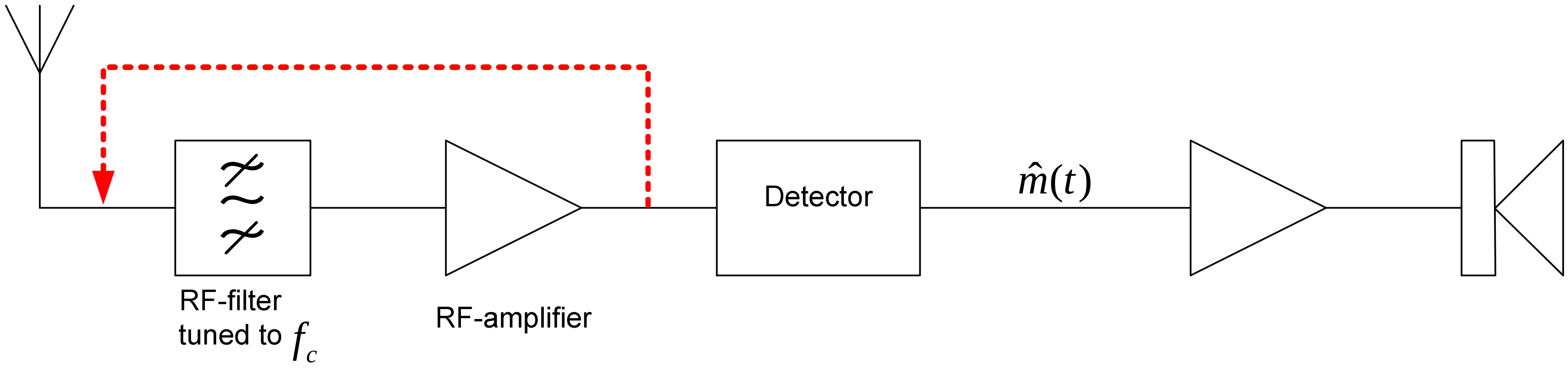
$$s(t) = \mathrm{Re}\{g(t)e^{j\omega_c t}\} + n(t) \qquad \hat{g}(t) = f(\hat{m}(t)) \qquad \hat{m}(t) = f^{-1}\{\hat{g}(t)\}$$



In the following, we discuss three types of receivers:

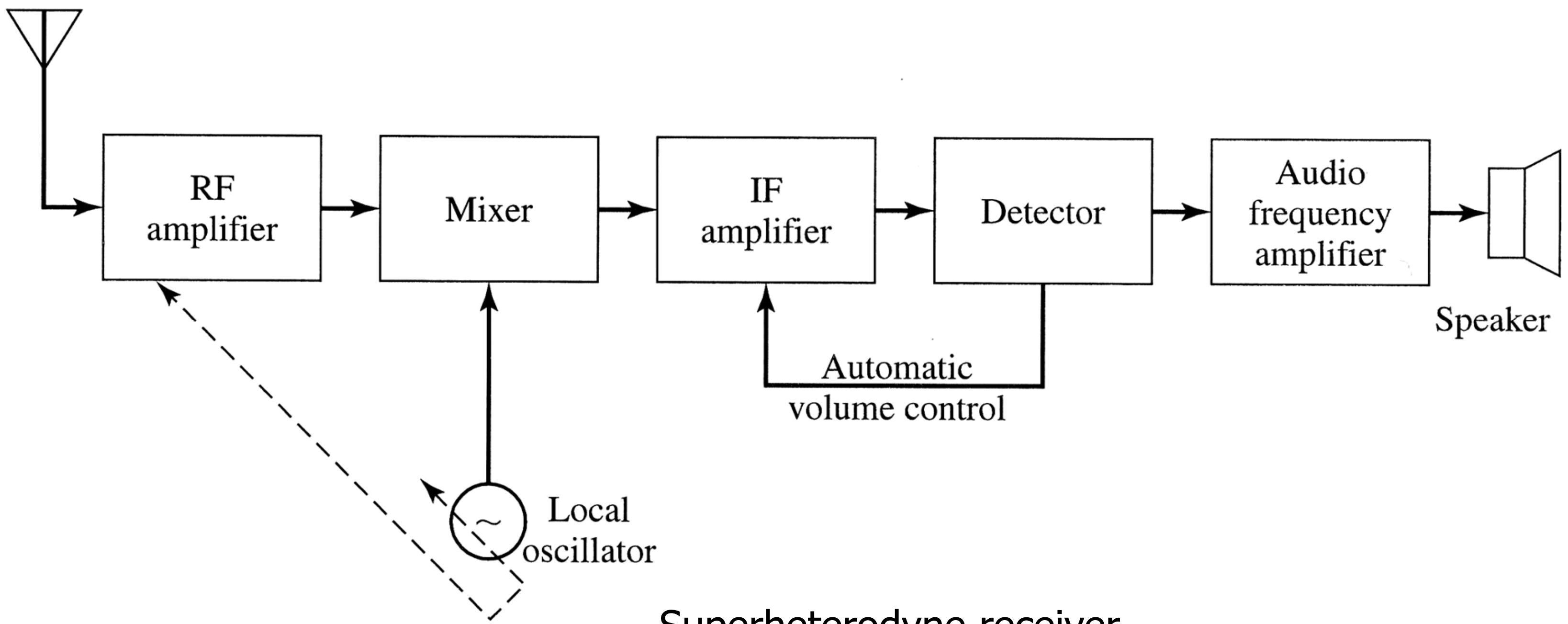
- the tuned radio frequency (RF) receiver
- the superheterodyne receiver
- the homodyne receiver

RF tuned receiver



Disadvantages:
- difficult to tune to a different frequency
- feed-back problems due to high amplification (oscillation)
- difficult to implement narrowband filters
  (sensitive to adjacent channel interference)

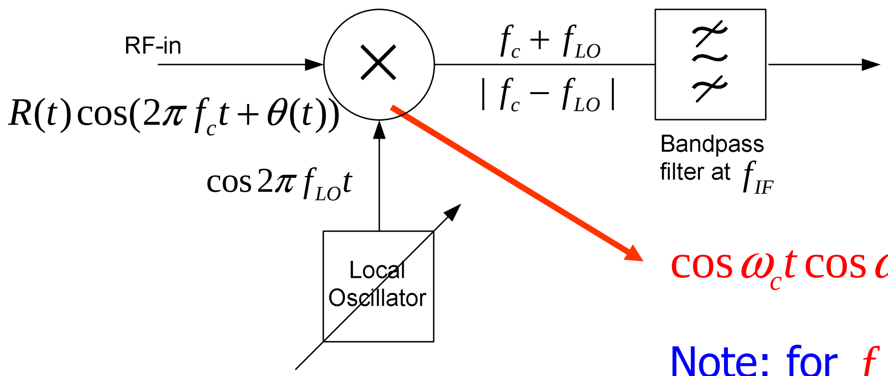
Superheterodyne receiver (1)



Superheterodyne receiver

How does it work?
The signal at the desired frequency is "mixed" to a fixed intermediate frequency (IF, "middenfrequentie") where all the processing: filtering, amplification and detection is done.

Superheterodyne receiver (2)

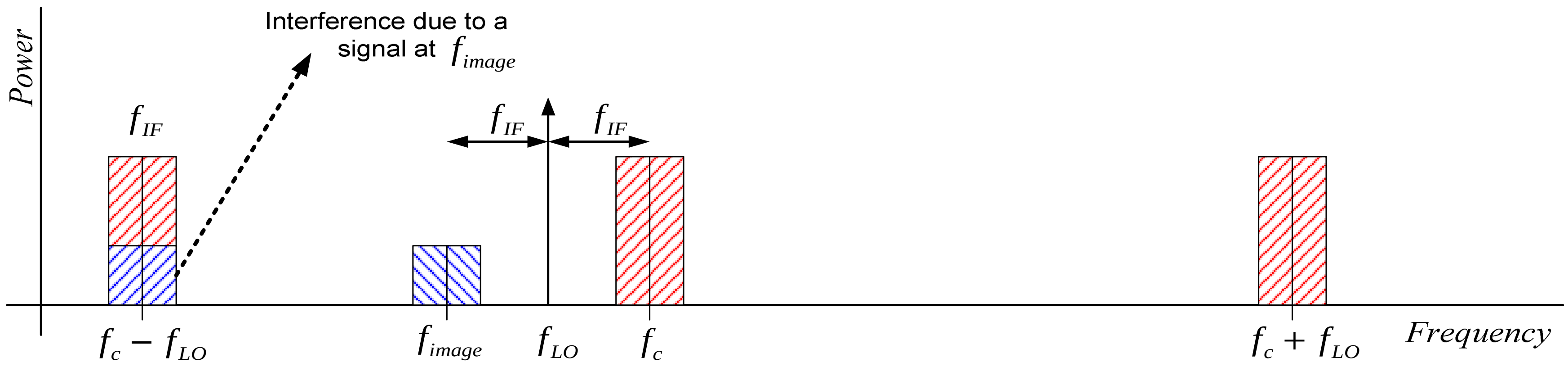


$f_{IF} = f_c + f_{LO} \rightarrow$ up-conversion

$f_{IF} = | f_c - f_{LO} | \rightarrow$ down-conversion

$$\cos\omega_c t \cos\omega_{LO} t = \tfrac{1}{2}[\cos(\omega_c - \omega_{LO})t + \cos(\omega_c + \omega_{LO})t]$$

Note: for $f_c < f_{LO}$ the sidebands are reversed (complex conjugate $g^*(t)$); this does not happen for $f_c > f_{LO}$.



Superheterodyne receiver (3)

What are image frequencies?

Image frequencies are frequencies (different from the desired frequency) that are also converted to the IF by the mixing operation.
Signals present at these frequencies will distort the reception of the signal at the desired frequency!

Image frequencies:

- down-conversion: $$f_{image} = \begin{cases} f_c + 2 f_{IF} & \text{if } f_{LO} > f_c \\ f_c - 2 f_{IF} & \text{if } f_{LO} < f_c \end{cases}$$ high-side down-conversion / low-side down-conversion

- up-conversion: $$f_{image} = f_c + 2 f_{LO} \quad \text{if } f_{IF} > f_{LO}$$ up-conversion

Signals at image frequencies have to be removed (attenuated) at RF-level, i.e. before the mixer!!!

Superheterodyne receiver (4)

Advantages of a superheterodyne receiver:

1. By using an intermediate frequency (IF)
   no feed-back problems, if $f_{IF} \neq f_c,\ f_{LO}$

2. Tuning by changing $f_{LO} \rightarrow$ a fixed or tunable input filter is needed
   to remove signals at the image frequencies.

3. With a well chosen IF, suitable signal amplification and filtering
   are possible.

Superheterodyne receiver (5)

Choice of IF:

Low IF-frequency:

1. Large amplification possible and cheap
2. Very narrowband filtering possible: $B \approx 0.001\, f_{IF} \;\; \rightarrow \;\; Q \approx 1000$

High IF-frequency:

3. Good suppression of image frequencies by RF-filtering simple and cheap

Example: AM-receiver

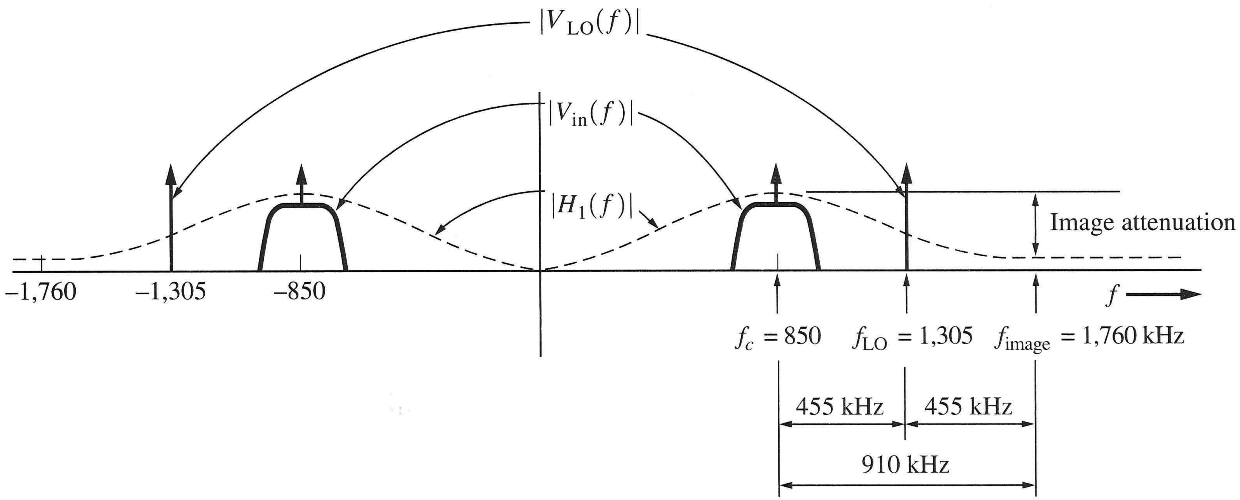


Figure 4–30 Spectra of signals and transfer function of an RF amplifier in a superheterodyne receiver.

The AM-receiver has an IF of 455 kHz and is tuned to $f_c = 850$ kHz. Since high-side down-conversion is applied ($f_{LO} > f_c$, $f_{IF} < f_{LO}$) the local oscillator frequency $f_{LO} = f_c + f_{IF} = 1305$ kHz.

The image frequency is found at $f_{image} = f_{LO} + f_{IF} = f_c + 2f_{IF} = 1760$ kHz, (check that $f_{image} - f_{LO} = f_{IF}$).

Super heterodyne receiver (6)

Standard IF-frequencies

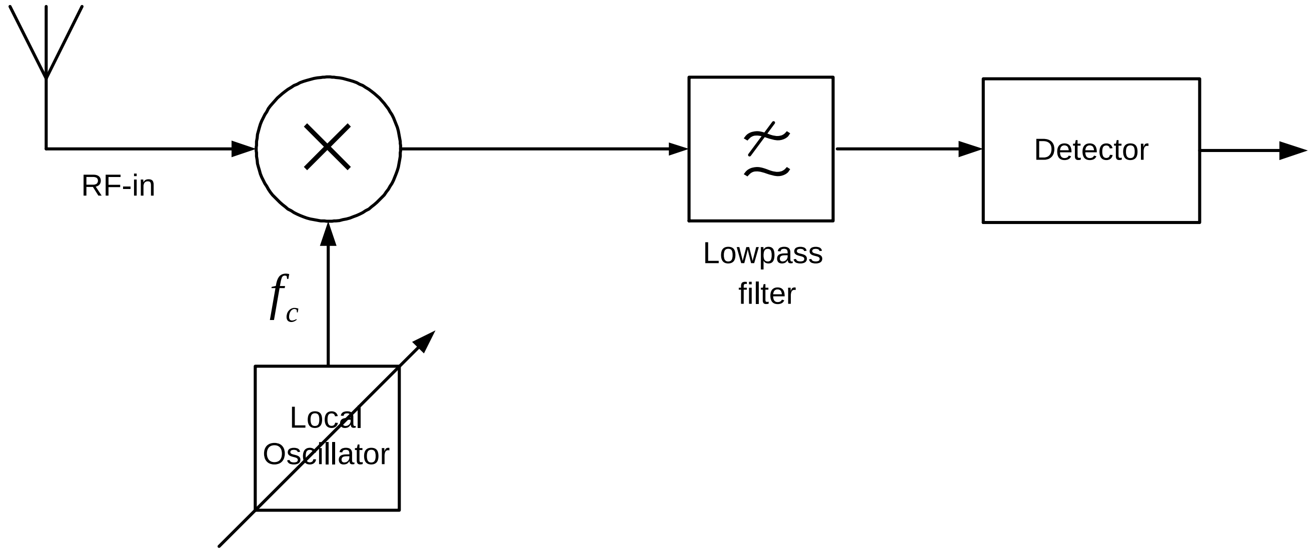
TABLE 4–4 SOME POPULAR IF FREQUENCIES IN THE UNITED STATES.

IF Frequency	Application
262.5 kHz	AM broadcast radios (in automobiles)
455 kHz	AM broadcast radios
10.7 MHz	FM broadcast radios
21.4 MHz	FM two-way radios
30 MHz	Radar receivers
43.75 MHz (video carrier)	TV sets
60 MHz	Radar receivers
70 MHz	Satellite receivers

Zero-IF receiver (1)

In a zero_IF receiver (homodyne detection or direct conversion), the signal is immediately converted to baseband:

- $f_{IF} = 0$
- $f_{LO} = f_c$



Zero-IF receiver (2)

Advantages of a zero_IF receiver:

1. No image frequencies
2. LO-frequency is equal to the frequency of the received signal
3. Broad range of applications and cheap
4. Via an AD-converter direct interface to a DSP
5. DSP software determines the receiver qualities
   $\longrightarrow$ software defined radio (SDR)

Disadvantages of a zero_IF receiver:

1. LO- signal leakage to the antenna
2. DC-offset after the mixer (e.g. due to LO leakage)
3. Low sensitivity: large and low-noise signal amplification needed.

Causes of interference in a receiver

1. Generated in the transmitter

   - signal power transmitted outside the allowed frequency band (spurious transmissions due to a non-stable oscillator)

   - higher harmonic signal components due to non-linearities

2. Generated in the receiver

   - non-linear distortion in the receiver front-end (amplifier, mixer), especially when strong input signals are present (spurious signals)

   - image frequencies, harmonic distortion in the LO

   - pass-through of signals outside the selected bandwidth due to lack of selectivity (Adjacent Channel Interference (ACI))