Modulation/Microwave Integrated Digital Wireless Developments

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Invited Paper

Abstract—**Simultaneous study and joint optimization of digital modulation techniques and of microwave components, presented in this paper, lead to spectrally efficient, high capacity, fast transmission, and throughput rate wireless systems. The choice of a particular modulation/demodulation (modem) technique has a major impact on the overall microwave system design, transceiver architecture and on the choice of all intermediate frequency (IF) as well as radio frequency (RF) component specifications. The performances of the most frequently used digital modulation techniques, including π/4-DQPSK, conventional BPSK, DQPSK, GMSK, GFSK, 4FSK, and FQPSK systems are compared. Study of crosscorrelated quadrature modulated GMSK systems used in GSM, PCS-1900, and other standards indicate that the subject technologies applied to GMSK improve the performance of these systems. We demonstrate that with the patented family of digitally modulated Feher's FQPSK systems, the power efficiency of conventional QPSK systems, which require linear microwave amplifiers, can be increased by about 300%, and the spectral efficiency of standardized nonlinearly amplified microwave integrated circuit GMSK systems can be increased by 60–200%. Similar advantages are obtained with our new FBPSK over BPSK modulated systems.**

I. INTRODUCTION

To meet the demand for the exploding capacity requirements by ever increasing numbers of users, powerful and spectrally efficient modulations must be implemented. New microwave and radio frequency (RF) technologies are being developed to fulfill this growing demand, not by designing more powerful transmitters and receivers, but by developing portable systems which are small, light, and have minimal power consumption requirements. In this paper, we review the most frequently used digital modulation formats along with newly developed techniques which lead to systems that are more power and spectrally efficient. In addition, some of the latest in microwave integrated circuit technologies are described. Tables of different wireless standards in the field of cellular, cordless telephone, and wireless local area network (WLAN) are presented in Appendix A at the end of this paper. The tables list different parameters including: Adopted modulation techniques, frequency bands, frequency range of operations, and transmission rate.

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Section II will review the principles and most important characteristics of modulation techniques which have been considered or are in use in various wireless, cellular, PCS and mobile standards and nonstandard large scale implementations. The most frequently considered modems include: BPSK, QPSK, Offset QPSK, π/4-DQPSK, FBPSK, FQPSK, GFSK, and GMSK. A unified presentation and simple overview of the principles of operation and resulting performance of each modem type is presented. Substantial parts of this paper bas been extracted from [1].

Section III will present some of the factors that have impacted the microwave component development in the field of wireless communications. They include size, battery life time, radiation, spectrum pollution, and robustness.. Evaluations of recently developed modulations using GaAs RFIC's, particularly in the 1.9 and 2.4-GHz bands, are also presented. Different applications and developments in the microwave and RF fields are highlighted.

II. REVIEW OF DIGITAL MODEM PRINCIPLES FOR WIRELESS SYSTEMS

The following digital modulation techniques, which are widely used in wireless communications, will be evaluated in both linear and nonlinear channels for their power and spectral efficiency. Bit error rate (BER) performance in both AWGN and Rayleigh-fading channels are also presented.

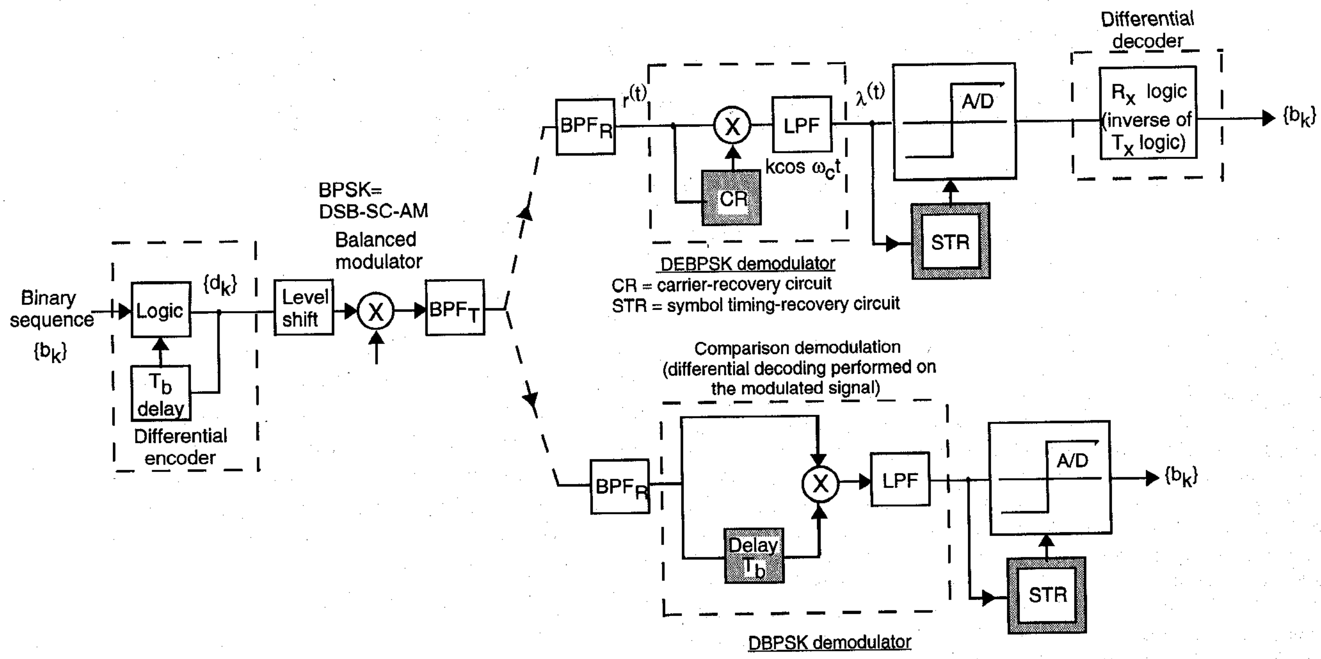
A. Coherent and Differentially Coherent Binary PSK (BPSK) Systems

Discrete phase modulation, known as M-ary phase-shift keying, is among the most frequently used digital modulation techniques [1]. Biphase or binary PSK systems (BPSK) are considered to be the simplest form of phase-shift keying ($M = 2$). Recently, it was adopted by the IEEE 802.11 Committee as the standard modulation in Direct Sequence Spread Spectrum (DS-SS) for WLAN (see Appendix A) [5]. A BPSK modulated signal has two states, $+1$, -1, as shown in Fig. 1(b). The BPSK modem is shown in Fig. 1(a). The modulated signal is given by

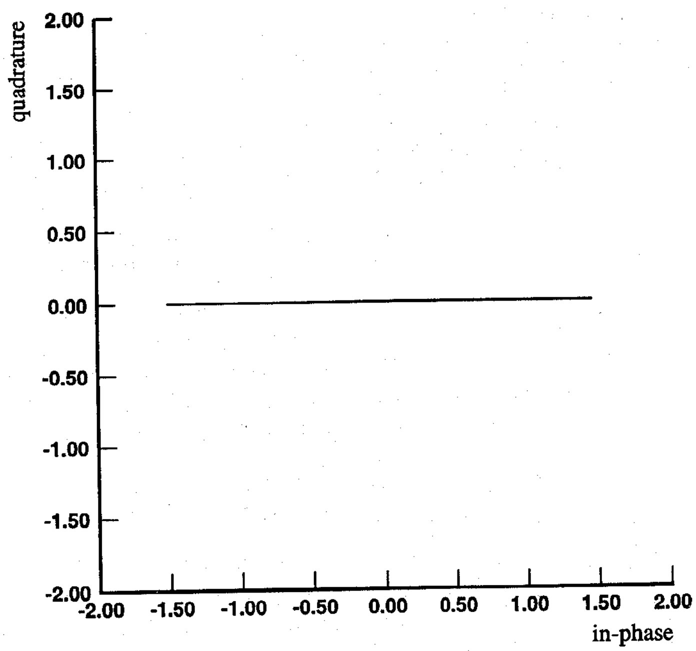
$$m(t) = b(t)c(t) = Cb(t)\cos\omega_c t \tag{1}$$

where $b(t)$ represents a synchronous random binary baseband signal having a bit rate $f_b = 1/T_b$ and levels -1 and $+1$, $\omega_c = 2\pi f_c$ is the carrier angular frequency, and C is the amplitude of the carrier signal (1), which can be written as

$$m(t) = C\cos[\omega_c t + \theta(t)]. \tag{2}$$



(a)



(b)

Fig. 1. (a) Differentially encoded BPSK modulator followed by coherent demodulator and differential decoder (DEBPSK) and differential phase demodulator (DBPSK). (b) Signal trajectory of a BPSK signal [1].

BPSK information is contained in the phase, $\theta(t)$, which takes the value of 0 or 180°. BPSK transmitted signals can be demodulated coherently and noncoherently. Coherent demodulation is implemented by multiplication of the modulated received bandlimited RF or IF signal, $r(t)$, by the recovered unmodulated carrier $k\cos\omega_c t$. Coherent phase modulation is removed in the carrier recovery subsystem. noncoherent modulation can be recovered using differential techniques. The differential BPSK demodulator performs a comparison detection (a demodulation) directly on the modulated signal, thus does not require a carrier-recovery circuit. The modulated signal is multiplied by a one-bit delayed replica and then bandlimited with a low-pass filter. The differentially encoded modem equipped with a carrier-recovery circuit is designated DEBPSK, while the one without carrier recovery is designated DBPSK [6].

B. QPSK-Quadrature Phase-Shift-Keying: Coincident and Offset QPSK (OQPSK)

Quadriphase or Quadrature-Phase-Shift-Keying (QPSK), and offset-keyed or staggered quadriphase modulation (OQPSK or SQPSK) modems are used in systems applications where the 1 bit/second/Hertz (b/s/Hz) theoretical spectral efficiency of BPSK modems is insufficient for the available

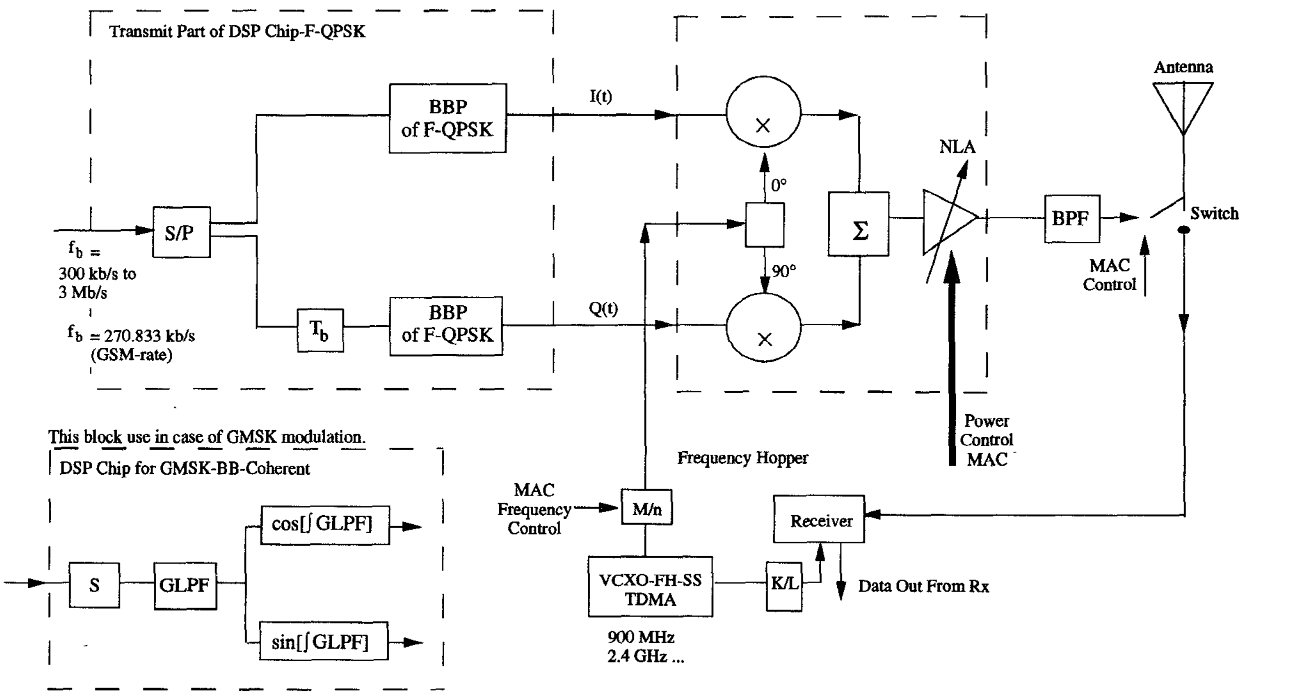


Fig. 2. Transmitter of a BB to RF radio for F-QPSK or GMSK nonlinear amplifier applications (slow frequency hopping-spread spectrum SFH-SS-TDMA).

bandwidth. The various demodulation techniques used in binary phase-shift-keyed systems also apply to quadriphase PSK systems. Quadriphase modulations are used in microwave radio, satellite applications, and were recently adopted for high data transmission of 2 Mbts/second for Direct Sequence Spread Spectrum WLAN applications [5].

A block diagram of a conventional, or "coincident," and of a staggered or offset QPSK modulator is illustrated in Fig. 2. For coincident QPSK applications, the $T_b = T_s/2$ offset delay line is removed (short circuited) from the quadrature $Q(t)$ channel. The NRZ data stream entering the modulator is converted by a serial-to-parallel converter into two separate NRZ streams, $I(t)$ and $Q(t)$, having a symbol rate equal to *half* that of the incoming bit rate. The data are later modulated using quadrature modulator structure, added, and amplified before transmission as shown in Fig. 2.

The four possible outputs of the coincident or conventional (not offset) QPSK modulator and their corresponding IQ digit combinations are shown in the signal space diagram of Fig. 3. Note that either 90 or 180° instantaneous phase transitions are possible. For an unfiltered QPSK signal, phase transitions occur instantaneously and the signal has a constant-amplitude envelope. However, phase changes for *filtered* QPSK signals result in a *varying envelope amplitude*. In particular, a 180° phase change results in a momentary change to zero in envelope amplitude. In OQPSK, the resulting instantaneous phase states at the modulator output are the same as for QPSK. However, because both data streams applied to the multipliers can never be in transition simultaneously due to the one bit delay in the Q channel, only one of the vectors that comprise the offset-keyed quadriphase modulator output signal can change at any one time. The result is that only 90° phase transitions occur in the modulator output signals. Similarly to QPSK, an *unfiltered* offset QPSK signal has a *constant-amplitude*

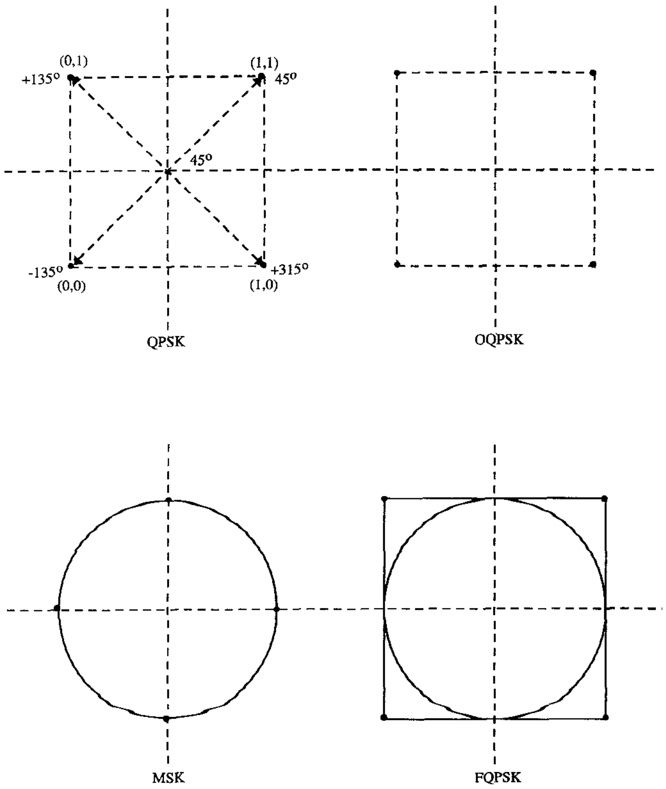


Fig. 3. Signal state transition trajectories for OQPSK, QPSK, MSK, and FQPSK.

envelope. However, for filtered offset QPSK signals the transmitted signal has an amplitude envelope variation of 3 dB (40%), compared to the 100% amplitude envelope variation for conventional QPSK systems. This lower-amplitude envelope variation imparts significant advantages to offset QPSK, as compared to QPSK in nonlinearly amplified power efficient cellular and other wireless system applications.

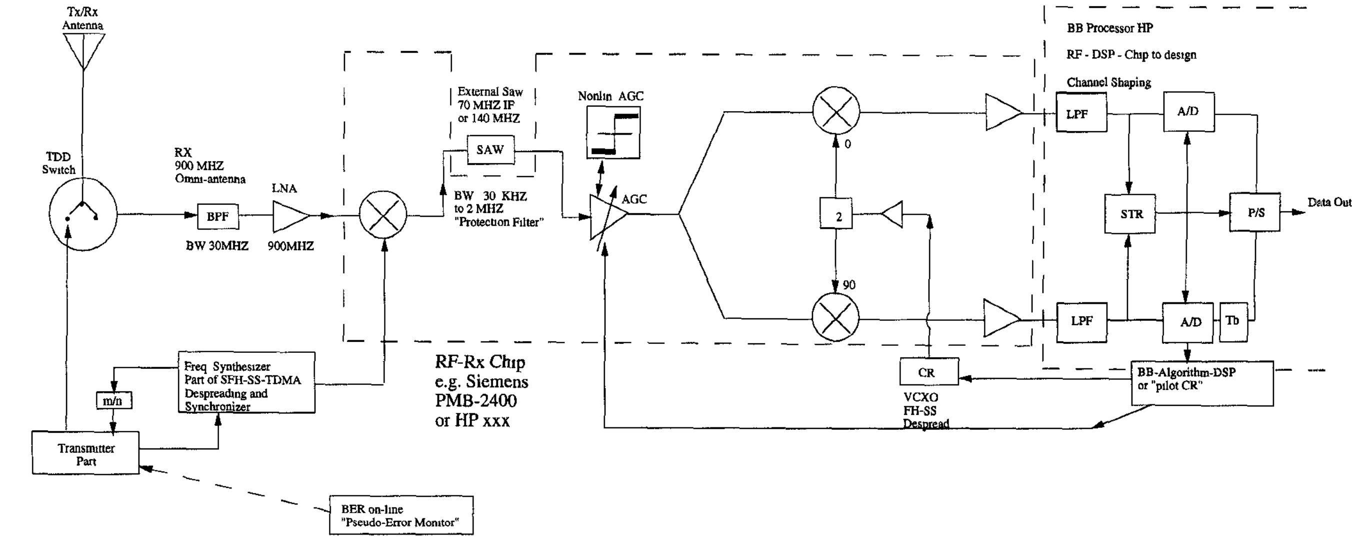


Fig. 4. Receiver with one IF stage for coherent F-QPSK or GMSK systems (TDMA-TDD-SFH-SS) [1].

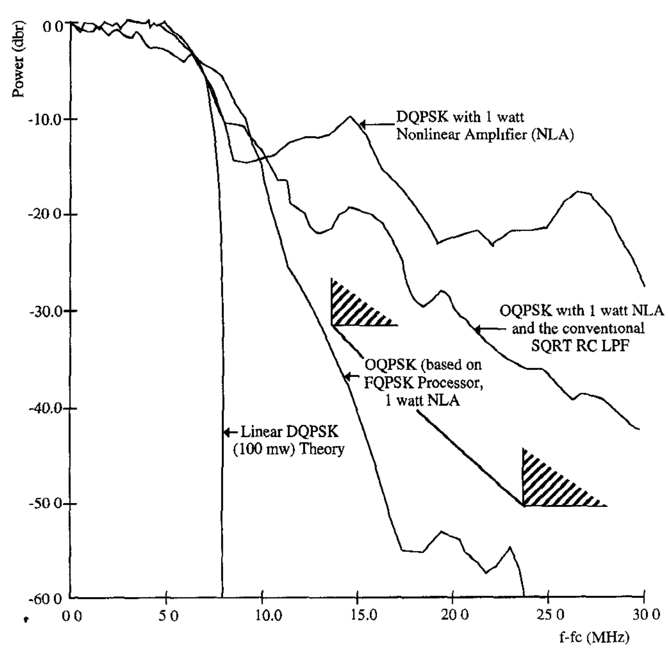


Fig. 5. Power spectral density in nonlinear amplified channel.

The QPSK signal at the modulator output is normally filtered to limit the radiated spectrum, amplified, and then transmitted over the transmission channel to the receiver input. Because the (I) and (Q) modulated signals are in quadrature (orthogonal), the receiver is able to demodulate and regenerate them independently of each other, operating effectively as two BPSK receivers. The regenerated (I) and (Q) streams are then recombined in a parallel-to-serial converter to form the original input data stream.

The offset O-QPSK receiver shown in Fig. 4 is identical to that for coincident QPSK with the exception that the regenerated (I) data stream is delayed by a unit bit duration $T_b = T_s/2$, so that when combined with the regenerated (Q) stream the original "input data" stream is recreated.

QPSK systems can be demodulated both coherently and noncoherently. Differential QPSK or DQPSK is used where simpler carrier recovery and fast synchronization is required at the expense of a maximum of 3 dB in performance degradation [6].

C. $\pi/4$-DQPSK Modems for US and Japanese Digital Cellular Standards

A modulation/demodulation (modem) method known as $\pi/4$ shifted, differentially encoded quadrature phase shift keying or $\pi/4$-DQPSK, has been adopted for US and Japanese digital cellular time division multiple access (TDMA) standards. This modulation method was introduced in 1962 by Baker of AT&T Bell Laboratories [7]. The $\pi/4$-DQPSK modulation technique represents a compromise solution between the conventional or coincident transition QPSK and OQPSK (offset QPSK) methods. Instantaneous phase transitions of

QPSK	are	$0°, \pm 90°, \pm 180°$
OQPSK	are	$0°, \pm 90°$
$\pi/4$-DQPSK	are	$0°, \pm 45°, \pm 135°$.

In power efficient, nonlinearly amplified (NLA), fully saturated, C-class or hard-limited cellular-mobile systems, the instantaneous 180° phase shift of conventional QPSK systems leads to a very significant spectral regrowth, and thus has a low spectral efficiency, see Fig. 5. In [5] and [9], it is demonstrated that conventional C-class amplifiers and hard-limited amplifiers or preamplifiers lead practically to the same spectral regrowth and BER performance degradation. In the case of $\pi/4$-DQPSK, the spectral restoration (regrowth) is somewhat reduced, as the instantaneous phase transitions are limited to $\pm 135°$.

Fig. 6 illustrates a conceptual implementation block diagram basic $\pi/4$-DQPSK modulator architecture.

The information is differentially encoded; symbols are transmitted as changes in phase rather than absolute phases. Let A_k and B_k denote the amplitudes of the unfiltered baseband non-return-to-zero (NRZ) pulses in the in-phase (I) and quadrature (Q) channel during $kT \leq t < (k+1)T$, respectively [8]. The signal levels A_k and B_k are determined by the signal levels

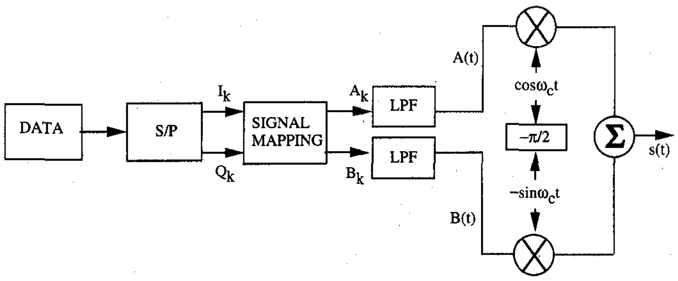


Fig. 6. π/4-DQPSK modulator block diagram.

TABLE I
PHASE CHANGE OF $\Delta\phi$ IN π/4-DQPSK SYSTEMS

I_k	Q_k	$\Delta\phi$
1	1	$\frac{-3\pi}{4}$
0	1	$\frac{3\pi}{4}$
0	0	$\frac{\pi}{4}$
1	0	$\frac{-\pi}{4}$

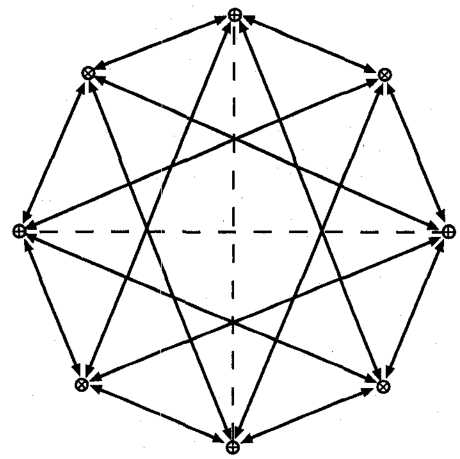


Fig. 7. Signal state constellations of a π/4-DQPSK system.

of the previous pulses and the current information symbol denoted by θ_k as follows:

$$A_k = A_{k-1}\cos\theta_k - B_{k-1}\sin\theta_k$$
$$B_k = A_{k-1}\sin\theta_k + B_{k-1}\cos\theta_k. \qquad (3)$$

In (3), θ_k is in turn determined by the current symbol of the carrier phase denoted by (I_k, Q_k) of the information source. The relationship between θ_k and the input symbol is given in Table I. A_k and B_k can take the amplitudes of ±1, 0, and $\pm 1/\sqrt{2}$. However, they are either two-level or three-level at any sampling instant. In Fig. 7, note the rotation by π/4 of the basic QPSK constellation for odd (denoted ⊕) and even (denoted ⊗) symbols [8].

The modulation architecture of π/4-DQPSK and π/4-QPSK systems is essentially the same as that of conventional DQPSK and of QPSK systems, described in previous sections, PI/4 DQPSK can be demodulated both coherently and noncoherently as described in details by Liu [8].

D. MSK (Minimum Shift Keying) Modems

Frequency Modulation (FM) is among the most frequently used analog modulation techniques. For data transmission a digital FM technique known as Frequency Shift Keying (FSK) was developed [10]. Logic state 1 corresponds to transmit frequency f_2, logic state 0 (−1 V data level) to f_1. The frequency deviation for coherent FSK is

$$\Delta f_{pp} = 2\Delta f = f_2 - f_1 = \frac{1}{2T_b} \qquad (4)$$

where T_b is the unit bit duration of the input data stream. The modulation index is defined by

$$m = \Delta f_{pp} \cdot T_b. \qquad (5)$$

The modulation index of FSK systems can be preset to have a narrowband or wideband digital FSK system. The FSK signal can be represented by

$$\begin{aligned} s_{\text{FSK}}(t) &= A\cos[2\pi(f_c \pm \Delta f)t] \\ &= A\cos(\pm 2\pi\Delta f t)\cos(2\pi f_c t) - A\sin(\pm 2\pi \Delta f t) \\ &\quad \cdot \sin(2\pi f_c t). \end{aligned} \qquad (6)$$

Simple noncoherent demodulators can be used for the demodulation of a large class of digital FSK signals [10]. Noncoherent demodulators could be simpler, yet require a higher C/N than coherently demodulated systems [1], [10].

Minimum Shift Keying (MSK) modulation is a special class of FSK with modulation index m equal to 0.5 [10]. It can be generated using a VCO structure as in FSK, or using a quadrature structure as premodulation filtered OQPSK as shown in Fig. 8. Modulation index of 0.5 value corresponds to frequency deviation $\Delta f = 1/4T_b$; thus the MSK signal $s_{\text{MSK}}(t)$ is

$$\begin{aligned} s_{\text{MSK}}(t) &= A\cos(\pm\pi t/2T_b)\cos(2\pi f_c t) \\ &\quad - A\sin(\pm\pi t/2T_b)\sin(2\pi f_c t). \end{aligned} \qquad (7)$$

Equation (7) is an MSK quadrature modulation representation of FSK. Note from Fig. 8, and (7) that the $\cos(\pm\pi t/2T_b)$ and $\sin(\pm\pi t/2T_s)$ represent **"independent, not correlated"** data streams after the serial to parallel converter The implementation method of the quadrature structure for an MSK modulator uses the same OQPSK modulator structure except the addition of sinusoidal pulse-shapers. In the baseband (I) channel, the pulse shaper generates a $\cos(\pm\pi t/2T_b)$ pulse sequence, while in the (Q) channel the offset delay, T_b, in conjunction with the pulse shaper, provides the pulse sequence given by

$$\cos\left[\pm\frac{\pi(t-T_b)}{2T_b}\right] = \sin\left(\pm\frac{\pi t}{2T_b}\right). \qquad (8)$$

The MSK demodulator operates in the same manner as the offset QPSK receiver described earlier. However, different filtering is required to ensure Inter-symbol Interference (ISI) free transmission [1].

In [1], the normalized power spectral density of conventional QPSK, offset QPSK, GMSK, and MSK as a function

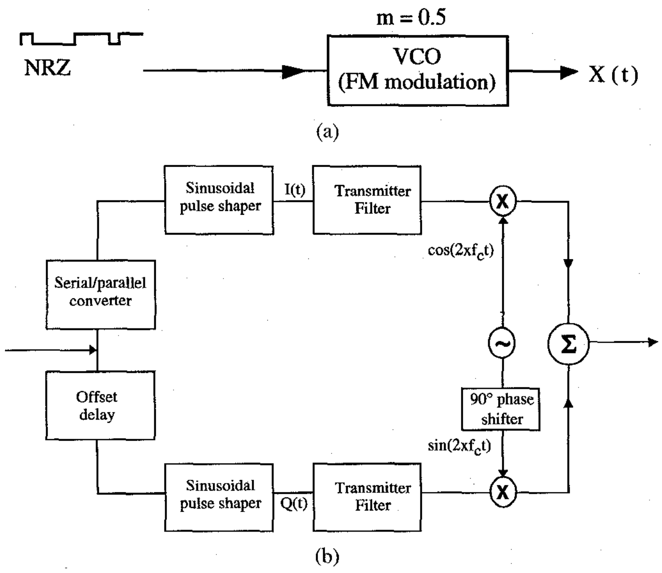


Fig. 8. MSK transmitter. (a) VCO structure; (b) quadrature structure.

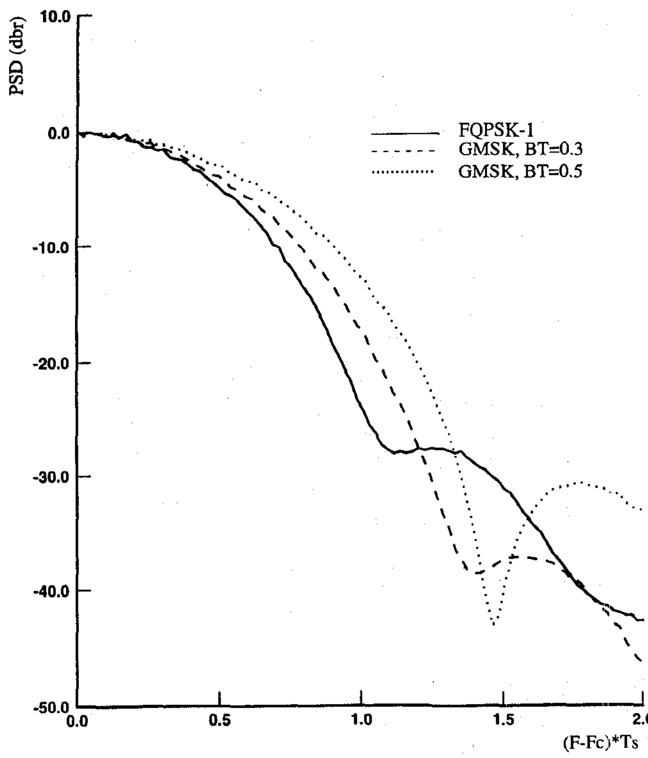


Fig. 9. PSD of FQPSK and GMSK in Nonlinear amplified channel.

of frequency, and normalized to the binary bit rate $R_b = 1/T_b$, have been evaluated. It was shown that the width of the main lobe of the MSK signal is $\pm 3/4T_b$, that is, the MSK signal has a 50% wider main lobe than the QPSK signal. For larger values of $(f - f_c)/f_b$, the unfiltered MSK spectrum falls off at a rate proportional to $(f)^{-4}$, the unfiltered QPSK spectrum at a rate proportional to $(f)^{-2}$. The unfiltered *infinite bandwidth* spectral properties are of particular importance in radio transmitter designs and applications where the output amplifier is frequently operated in a nonlinear (saturated) power efficient mode.

E. GMSK and GFSK

MSK modulation described in the subsection has the following fundamental properties:

1) Constant envelope suitable for nonlinear power efficient amplification
2) Coherent and noncoherent detection capability
3) Spectral main lobe is 50% wider than that of QPSK signals. First spectral null is at $(f - f_c)T_b = 0.75$ instead of $(f - f_c)T_b = 0.5$.

To retain the desirable 1) and 2) properties and to simultaneously increase the spectral efficiency, or reduce the bandwidth of the main lobe and the spectral density of the sidelobes, a Gaussian Premodulation Lowpass Filter (GLPF) is inserted into the baseband processor (BBP) subsystem of the MSK modulator. The cascade of a G-LPF with an FM modulator—VCO with $m = 0.5$, leads to a Gaussian-MSK or GMSK modulator [17], [18].

The pulse response of Gaussian Low-Pass Filters (GLPF) has a direct impact on both spectrum and performance of GMSK systems. This response is determined by the BT_b product where B is the 3 dB bandwidth of the Gaussian filter and T is the bit duration [9]. $BT_b = \infty$ corresponds to MSK system. A smaller BT_b leads to more compact spectrum but at the same time more ISI. Hence, the choice of BT is a compromise between spectrum efficiency and BER performance.

1) VCO-FM Based Implementation of GFSK and GMSK Systems: The VCO-FM modulated output is given by

$$x(t) = \cos\left[2\pi f_c t + \phi(t)\right] \quad (9)$$

where

$$\phi(t) = k \int_{-\infty}^{t} b(v)\, dv = k \int_{-\infty}^{t} \sum_{n=-\infty}^{\infty} a_n r(v - nT)\, dv \quad (10)$$

and k is a proportionality constant directly related to the sensitivity of the VCO-FM modulator. In Gaussian prefiltered FSK systems (GFSK), the modulation index m can be chosen in a wide range, for example $0.1 \leq m \leq 1$ in several wireless applications [11]. In GMSK systems, m must be preset to an exact value of 0.5, see (5). In conventional GMSK, k is chosen such that the contribution of a single phase pulse $r(t)$ to the change of the modulated phase $\phi(t)$ is exactly $\pi/2$

$$k \int_{-\infty}^{\infty} r(t)\, dt = \pi/2 \quad (11)$$

where $r(t) = \Pi(t/T) \times g(t), g(t)$ is the impulse response of GLPF, and $\Pi(t)$ is a rectangular pulse with a unit amplitude defined between 0 and T_b. The modulated transmitted signal $x(t)$ is a constant envelope compact spectrum signal [12], [7].

2) Quadrature GMSK—an Improved Implementation: The MSK and GMSK modulated/radio architecture, based on a baseband subsystem followed by an IF or direct RF VCO-FM radio transmitter, is simple (see Fig. 10). However, it may not be suitable for coherent demodulation. For coherent demodulation, the modulation index m must be exactly $m = 0.5$ [11]. The modulation index m of conventional VCO and VCXO (Voltage Controlled Crystal Oscillators) transmitters drifts over time and temperature.

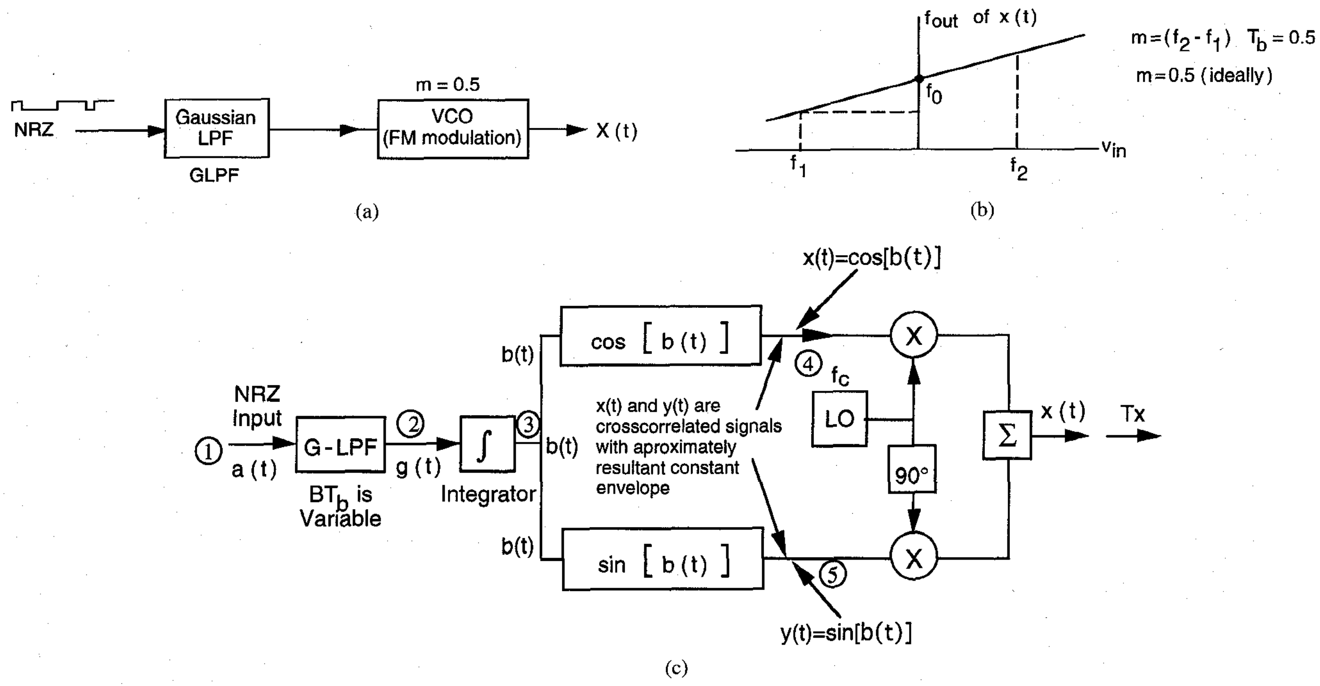


Fig. 10. GFSK and GMSK implementation/architectures. (a) VCO implementation. (b) Transfer function of a voltage controlled oscillator (VCO) based binary FM modulator. (c) A frequently used "crosscorrelated signal processor" architecture implementation of GMSK [1].

An alternative GMSK implementation architecture has a quadrature baseband processor (BBP), followed by a quadrature (QUAD) modulator. See Fig. 2. In this implementation the modulation index is exactly $m = 0.5$. This QUAD architecture leads to the same GMSK modulated signal as the VCO-FM based implementation as shown below.

The VCO-FM modulated output $x(t)$

$$\begin{aligned} x(t) &= \cos\left[2\pi f_c t + \phi(t)\right] = \cos 2\pi f_c t \cos\phi(t) \\ &\quad - \sin 2\pi f_c t \sin\phi(t) \\ &= \cos 2\pi f_c t \cos\left[k \int_{-\infty}^{t} b(v)\, dv\right] - \sin 2\pi f_c t \\ &\quad \cdot \sin\left[k \int_{-\infty}^{t} b(v)\, dv\right]. \end{aligned} \tag{12}$$

The quadrature structure implements the last part (right hand side) of (12). From (10), we have $\phi(t) = k \int_{-\infty}^{t} b(v)\, dv$, where $b(t)$ is the output of the G-LPF. Note that in (12) the I and Q baseband signals $b(v)$ are the same signals in GMSK which is contrary to the MSK implementation of (7), and Fig. 8.

Numerous noncoherent GMSK implementations have been investigated and described in the literature [13]. The conventional FM limiter/discriminator has been widely implemented by the industry for demodulating MSK/GMSK signals. The performance of the discriminator detector is not very sensitive to the inaccuracy of the modulation index which is an advantage for simple discriminator detectors [11].

The coherently demodulated baseband signals $\cos[\phi(t)]$ and $\sin[\phi(t)]$ contain considerably less ISI than the premodulation LPF-d transmitted signal. For $B_bT = 0.25$, the transmitted eye diagram has 3 and 6 levels in alternate sampling instants resulting in a smaller eye opening, while the corresponding quadrature demodulated (I) and (Q) channel eye diagrams each have two levels with some ISI but wider eye opening as shown in Fig. 11 [13].

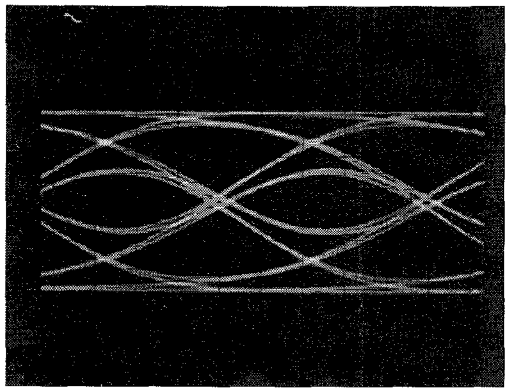


Fig. 11. Received eye diagram of noncoherent demodulated GMSK signal.

The ideal BER $= f(E_b/N_o)$ performance of coherent systems is better than that of noncoherent GMSK systems [13], [18].

3) GMSK Quadrature Implementation: The crosscorrelated quadrature modulated GMSK implementation, illustrated in Fig. 10(c) [4] is used in numerous recently released IC chips manufactured to the Global Mobile System (GSM) standard [4], [6]. In this section, we demonstrate that the implementation requires using crosscorrelation in the transmitter. Quadrature crosscorrelated modulation techniques have been invented and patented by Kato/Feher (for short "KF") [4]. Digital communications textbooks and prestigious research publications contain detailed justifications in regard to the need of having independent or uncorrelated in-phase (I) and quadrature (Q) baseband signals. Independent I and Q signals lead to optimum quadrature modulator performance. Cross-correlation between the I and Q baseband transmit signals would be a cause for crosstalk and thus detrimental

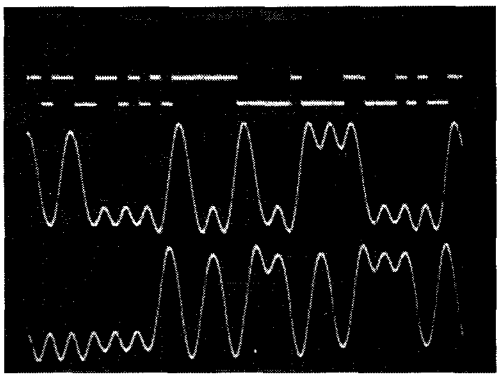


Fig. 12. GMSK baseband signal with $BT_b = 0.3$. Note the correlation between I, Q signals.

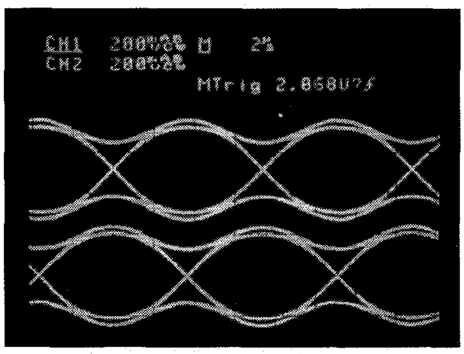


Fig. 13. Eye diagram of coherently demodulated GMSK signal.

to performance. Therefore, communication theory developed for linear systems, demonstrates that cross-correlation is not desired and that for optimum performance it must not exist.

The Kato/Feher correlated signal processor patent demonstrates, contrary to previous theoretical and practical achievements, that correlation between the I and Q channels can be of substantial benefit [1], [3], [32].

4) Correlated and Crosscorrelated In-Phase (I) *and Quadrature* (Q) *Signals of GMSK Transmitters:* An illustration of the fact that the in-phase (I) and Quadrature (Q) signals in a GMSK transmitter are crosscorrelated [3] follows. In Fig. 10(c), we refer to the input NRZ signal $a(t)$ at point 1, Gaussian low-pass filtered signal $g(t)$ at point 2, integrated and Gaussian filtered signal $b(t)$ at point 3, and cosine and sine values of the $b(t)$ signal, designated as $x(t)$ and $y(t)$ at points 4 and 5 of Fig. 10(c). Signals for the implementation technique of Fig. 10(c) have been generated for the **Global Mobile System (GSM) international standard specifications.** These specifications stipulate a GMSK modulator with $BT_b = 0.3$. A brief explanation of the generated signals of Fig. 10(c), and of the corresponding eye diagrams follows.

The Gaussian ($BT_b = 0.3$) filtered output signal at point 2 in Fig. 10(c) of an input Non-return to Zero (NRZ) signal for a typical random pattern is illustrated in Fig. 12. Note that this Gaussian filtered NRZ signal pattern, designated $g(t)$, exhibits substantial Intersymbol Interference (ISI). The ISI-free maximal amplitude values are normalized to $+1$ and -1. After integration, the filtered and integrated signal occurs at point 3. This random-like data signal, signal $b(t)$, is not a "sinusoidal" signal. At point 3, this signal is split into the "in-phase" (I) and "quadrature" (Q) channels. Thus the **same** $b(t)$ signal appears in the I and Q channels. The presence of the "same" I and "Q" is a fundamental difference between MSK, OQPSK, QPSK, and GMSK quadrature implementations in which a serial to parallel converter provides independent and **not correlated** binary data into the I and Q branch of the modulator. Thus splitter at point 3, with its connecting wires, provides the same $b(t)$ signal into I and Q. Evidently these I and Q signals are **100% correlated and crosscorrelated** as they are the same. An interesting implementation technique (to **split** the **same** signal into I and Q).

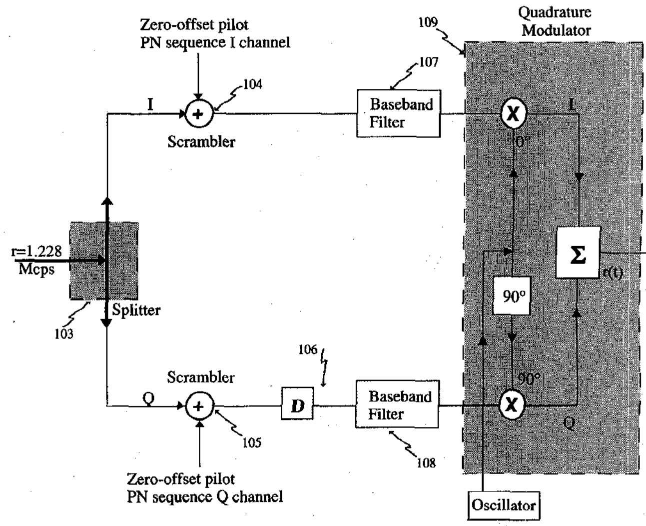


Fig. 14. CDMA reverse link structure as specified in the IS-95 standardized CDMA specifications [1].

Following the $\cos[b(t)]$ and $\sin[b(t)]$ processors at points 4 and 5, the generated signal patterns are designated $x(t)$ and $y(t)$. Points 4 and 5 serve as the in-phase and quadrature drives of quadrature modulators. These $x(t)$ and $y(t)$ random data signal patterns are "nonsinusoidal" in nature, i.e., they are not periodic sinusoidal waves. The $x(t)$ and $y(t)$ signals are crosscorrelated, in fact, they are related by equations.

$$x(t) = \cos[b(t)] \tag{13a}$$

$$y(t) = \sin[b(t)]. \tag{13b}$$

Thus, the "predictability" or crosscorrelation of $y(t)$ from $x(t)$ is defined. In this case,

$$y(t) = \sin[\cos^{-1}x(t)]. \tag{14}$$

In other words, to generate $x(t)$ and $y(t)$ we use a mathematical relation of "crosscorrelation" between these terms.

In Fig. 13, the respective "eye diagrams" of these crosscorrelated $x(t)$ and $y(t)$ signals are displayed. Note that these crosscorrelated signals are data transition jitter free and the maximum magnitude of the quadrature (alternatively in-phase) signal is reduced when the in-phase channel signal is non-zero. This leads one to conclude that the GMSK system quadrature implementation is using crosscorrelation.

5) Offset QPSK (OQPSK) Modulation as Used in the CDMA Standard IS-95: In Fig. 14, the modulator structure as standardized by the IS-95 committee for CDMA spread spectrum systems for the reverse link is illustrated. It is worthwhile to note that the specified OQPSK system is a crosscorrelated quadrature modulated spread spectrum system. The crosscorrelation between the I and Q channel is introduced by the splitter as well as the scrambler and baseband filters. In conventional OQPSK systems, instead of splitter which provides the same data into the I and Q channel, a serial to parallel converter is used. It is interesting to note the structural similarities between FBPSK, GMSK, and crosscorrelated OQPSK structure of the CDMA reverse link [1], [3].

F. FQPSK Family of Increased Capacity Modem/Radio

p1) FQPSK: A Superior Modulation for PCS and Mobile Cellular Radio: A family of offset-QPSK modems combined with power efficient nonlinear radio amplifiers known as "FQPSK" is described in this section. This new family of modulations is being developed to improve the power and spectral efficiency over current modulation techniques. FQPSK has a 50% higher spectral efficiency and a superior BER $= f(\mathrm{C/N})$ performance than the Digital European Cordless Telephone (DECT) and other standardized GMSK systems [2], [14], [15]. The hardware simplicity constant envelope and suitability of nonlinear power amplification of FQPSK also make it superior to the π/4-DQPSK standardized systems in North America and Japan. This combined spectral efficiency and BER performance advantage could nearly double (95% increase) the capacity of the DECT system [16].

The FQPSK family of modems/radio systems may operate with fully saturated C-class and/or hardlimited amplifiers, while the π/4-DQPSK radio systems could require a 4–7-dB-output backoff. These output backoff (OBO) requirements are introduced by the following practical constraints.

The block diagram of a generic FQPSK modulator, including a nonlinear RF power amplifier, is illustrated in Fig. 2. Where the FQPSK processor is inserted in both (I) and (Q) channels and a delay of one bit is inserted in the (Q) channel, additional filtering following the processor can be added for further improvement of spectral efficiency. At the output of the low power hard limiter, the FQPSK signal is a constant envelope. This hard-limiter could be implemented at an IF frequency or directly at RF.

The FQPSK pulse shaping filter has an impulse response $p(t)$ described by

$$p(t) = \begin{cases} 0.5\left[1+\cos\left(\dfrac{\pi t}{T_s}\right)\right] & \text{for } |t| \le T_s \\ 0 \end{cases} \tag{15}$$

where T_s is symbol duration. After the quadrature modulator and hardlimiter, we obtain the FQPSK signal $S_o(t)$

$$S_o(t) = \frac{x(t)\cos 2\pi f_c t}{\sqrt{x^2(t)+y^2(t)}} + \frac{y(t)\sin 2\pi f_c t}{\sqrt{x^2(t)+y^2(t)}} \tag{16}$$

where

$$x(t) = a_n p(t - nT_s) + a_{n-1} p[t-(n-1)T_s]$$

and

$$\begin{aligned} y(t) =& b_n p[t-0.5](T_s) + b_{n-1} p[t-(n+0.5)T_s] \\ &+ b_{n-2} p[t-(n+1.5)T_s] \end{aligned} \tag{17}$$

are the (I) and (Q) channel baseband signals, respectively.

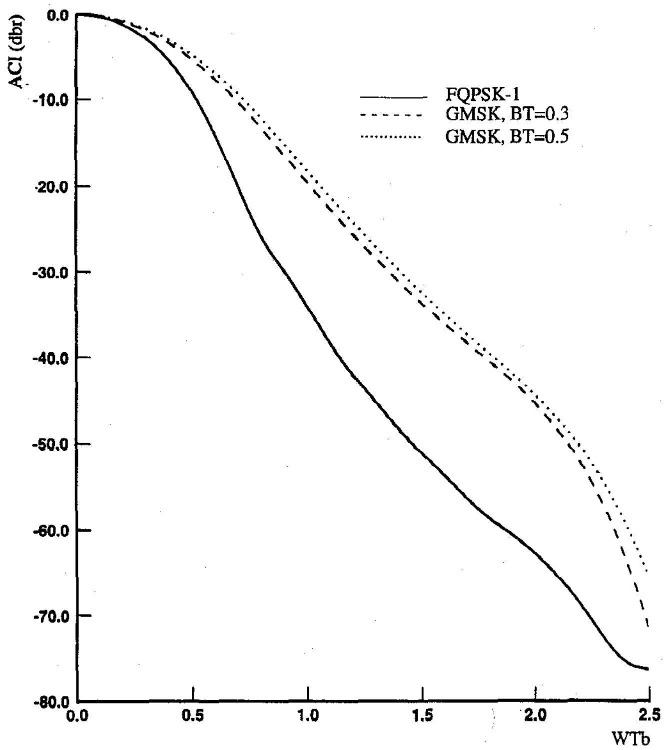


Fig. 15. Adjacent Channel Interference (ACI) of FQPSK and GMSK.

Radio and modem transmitter/receiver diagrams suitable for FQPSK, conventional QPSK, MSK, and GMSK implementations are illustrated in Figs. 2 and 4. In this case, a Slow Frequency Hopping (SFH)—Spread Spectrum (SS) configuration is illustrated. A direct baseband processor (BBP) to radio frequency modulation architecture is depicted. The transmitter implementation is practically the same for GMSK and FQPSK systems. In the FQPSK baseband processor (BBP) implementation of $p(t)$, (15) is somewhat simpler than that of GMSK systems.

Coherent FQPSK and GMSK receivers are very similar, see Fig. 4. The overall optimal channel shaping filter (receiver BPF in cascade with the equivalent post demodulation lowpass filters) is for

GMSK	Gaussian	with $BT_b = 0.6$	[18]
FQPSK	Butterworth	with $BT_b = 0.55$	[16]

Fig. 9 illustrates the power spectral density (PSD) of nonlinearly amplified (hardlimited or C-class amplified) FQPSK and GMSK systems. The out-of-band integrated Adjacent Channel Interference (ACI) of these nonlinearly amplified systems is shown in Fig. 15. The computed ACI ratio is defined by

$$\text{ACI ratio} = A(W) = \frac{\displaystyle\int_{-\infty}^{\infty} G(f)|H(f-W)|^2\,df}{\displaystyle\int_{-\infty}^{\infty} G(f)|H(f)|^2\,df} \tag{18}$$

where W is the normalized channel spacing, $G(f)$ is the power spectral density of the modulated/nonlinearly amplified signal, and $H(f)$ is the transfer function of the receiver channel

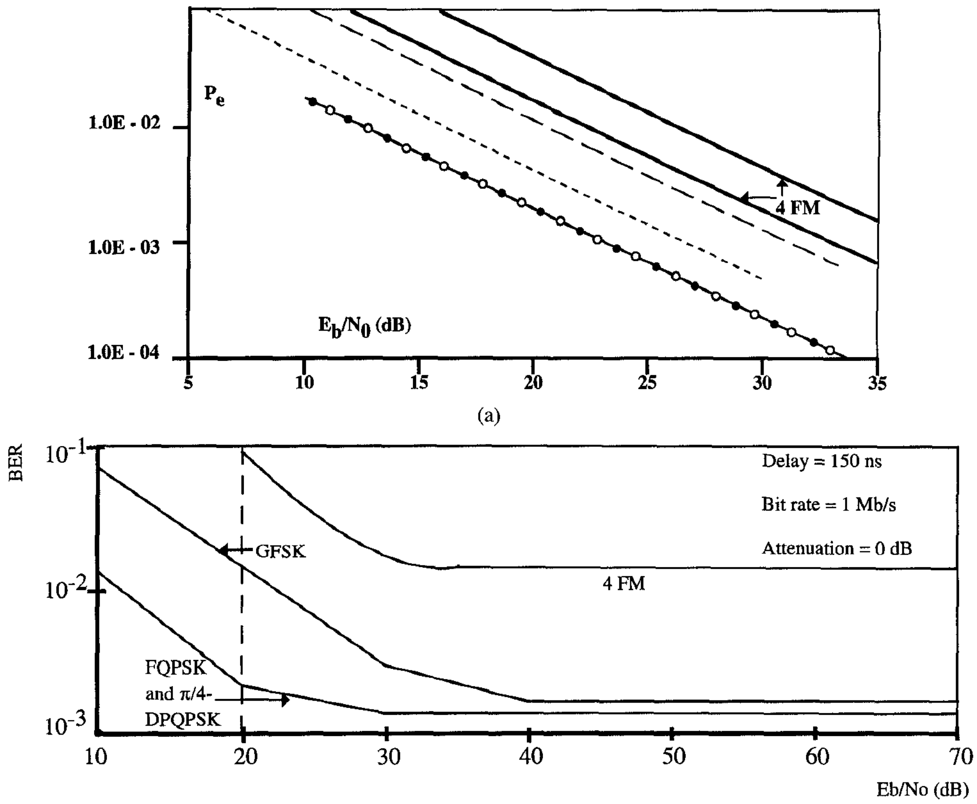


Fig. 16. BER $= f(E_b/N_o)$ performance of FQPSK, GMSK, and 4-FM constant envelope systems in (a) Rayleigh and (b) Rayleigh/delay spread environment. •—•—•, QPSK (lin. Ch.); — — — , GMSK noncoherent; o—o—o, F-QPSK (H/L Amp.); - - - - - -, GMSK BT-0.3. coher.

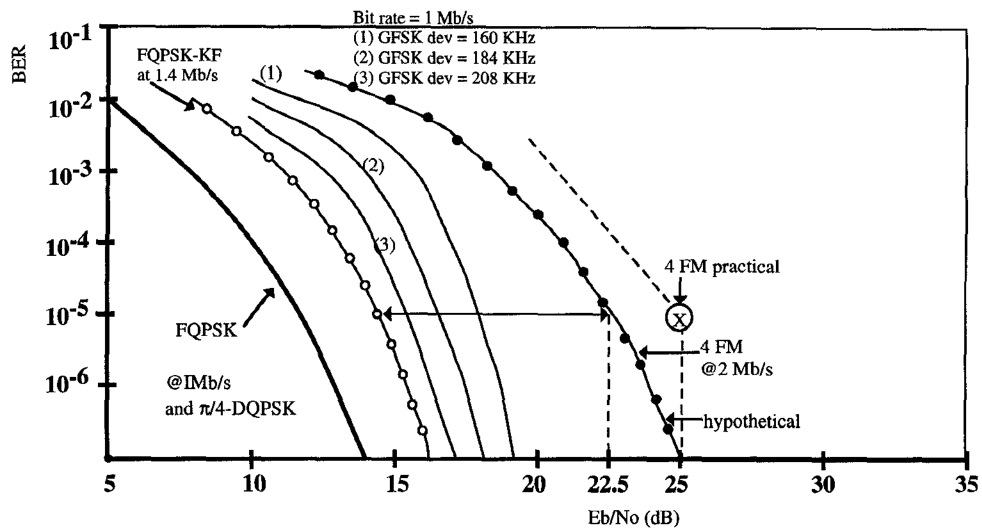


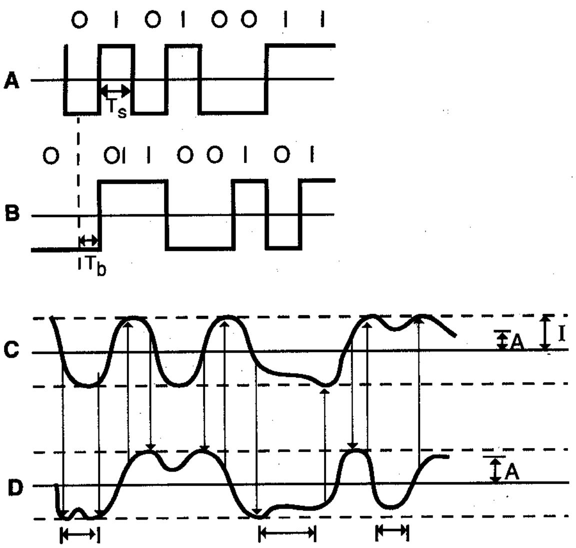
Fig. 17. Performance comparison in a stationary Gaussian noise environment [1].

shaping filter (cascade of the band-pass and of the demodulated low-pass) filters. Various standardization committees have somewhat different definitions of integrated ACI.

In practical system optimizations of GMSK and of FQPSK radio systems, four classes of cascaded bandpass equivalent lowpass filters have been considered [16] and [18]:

In Fig. 15, the computed integrated ACI ratio, $A(\omega)$, as a function of normalized channel spacing, WT_b, for FQPSK and GMSK systems is illustrated. From the results we note that FQPSK causes far less ACI than GMSK for $WT_b < 1.5$. To obtain this result, we employ a 4th order Butterworth BPF with normalized 3 dB bandwidth $B_iT_b = 0.55$ for FQPSK and a 4th order Gaussian BPF with $B_iT_b = 0.6$ for GMSK. From the ACI graphs we note that FQPSK has a 50% spectral advantage over standardized GMSK systems with $B_iT_b = 0.5$ and 0.3. These filters are chosen for their optimized BER performances in Gaussian noise channels.

The BER $= f(E_b/N_o)$ performance curves [15], [16] in stationary Additive White Gaussian (AWGN) and in mobile slow Rayleigh faded channels are highlighted in Figs. 16 and 17, respectively. In these graphs the performance of ideal theoretical, linearly amplified QPSK, and of practical nonlinearly



Signal after cross correlation

Fig. 18. FQPSK-KF baseband waveforms [1], [3].

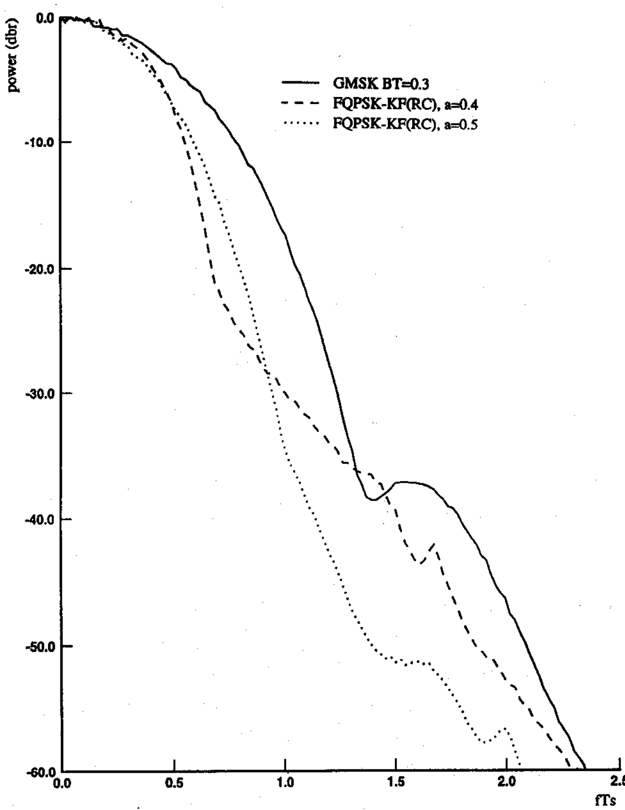


Fig. 19. PSD of FQPSK-KF in nonlinearly amplified channel (NLA).

amplified FQPSK and GMSK modems are presented. From Fig. 17, we note that the performance of FQPSK is practically equal to the linearly amplified QPSK, while the performance of GMSK is about 2.5 dB worse than that of QPSK.

2) FQPSK-KF: Another member of the FQPSK family is "FQPSK-KF" [3], [19] which led to even more significant performance improvements than the first FQPSK technique described in the previous subsection. A detailed description of FQPSK-KF, previously known as XPSK or cross-correlated PSK, is given in [3], [19], [20].

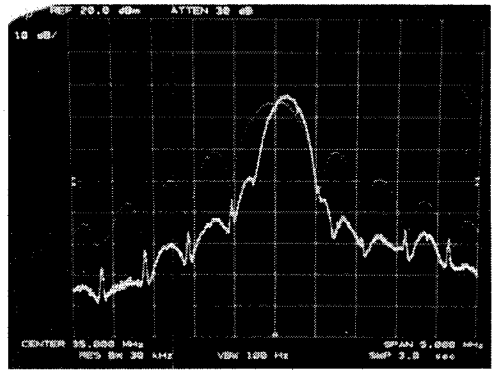
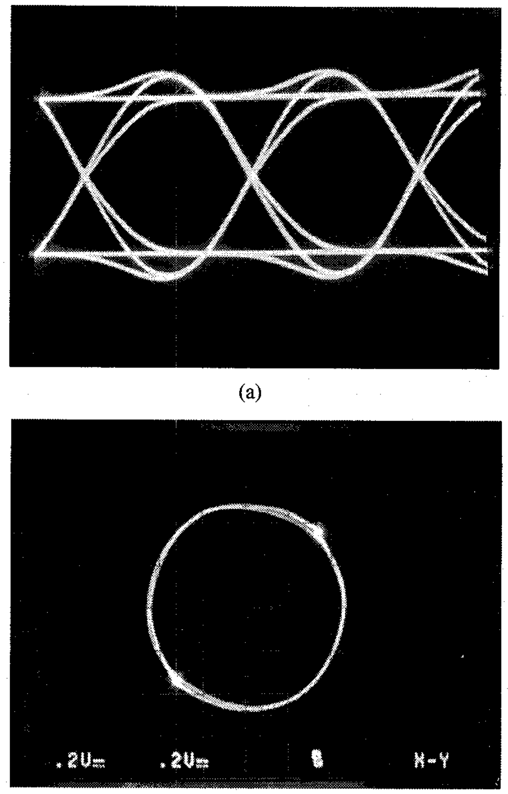


Fig. 20. PSD of BPSK (upper trace) and FBPSK (lower trace) in nonlinear channels.



(a)

(b)

Fig. 21. (a) An eye diagram of a FBPSK signal, (b) signal trajectory of a FBPSK signal.

One of the most important concepts in this technique is the cross-correlation between the in-phase and offset quadrature phase filtered signals. A cross-correlator of Intersymbol and Jitter Free (IJF) baseband signals and the corresponding baseband signal drives of the offset QPSK modulator are illustrated in Fig. 18. The correlation between I and Q is based on a correlator factor, a, that ranges between 0.5 and 1. For $a = 1$, there is no correlation between I and Q and the original FQPSK signal is obtained. The nonlinearly amplified power spectral density is shown in Fig. 19. A spectral efficiency comparison is presented in Table II. In [21]–[25] it is demonstrated that FQPSK-K offers up to 300% capacity increase over the IEEE 802.11 GFSK standardized system.

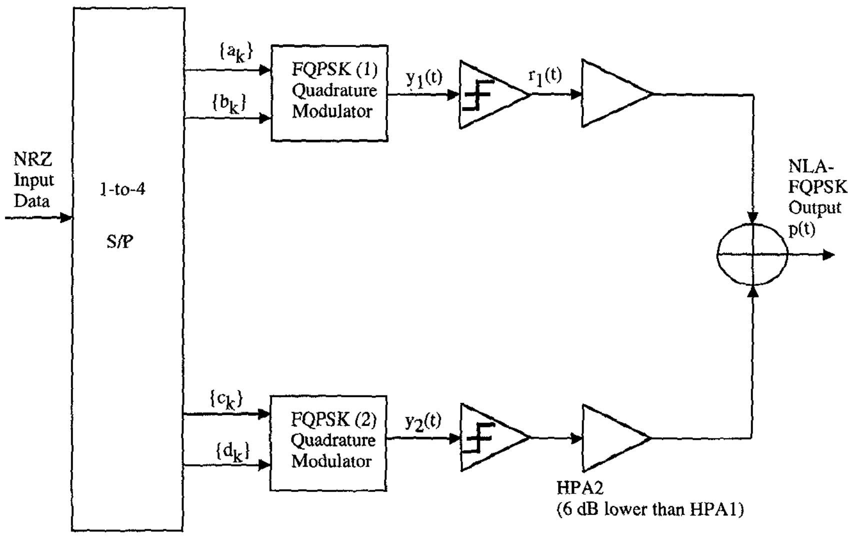


Fig. 22. FQPSK4 × 4 modulator structure.

TABLE II
SPECTRAL EFFICIENCY COMPARISON OF NONLINEARLY AMPLIFIED GMSK, GFSK, FQPSK, AND FQPSK-KF BASED ON FCC PART 15.247 DEFINITION OF 99% IN-BAND POWER [3], [4], [11]

GMSK	GFSK	FQPSK	FQPSK-KF
0.8	1	1.22	1.5

3) FBPSK: FBPSK is another member of the FQPSK family which leads to nonlinearly amplified power efficient BPSK modems. This modulation is based on using a quadrature modulator structure as in QPSK systems [26]. The FBPSK block diagram is the same as an FQPSK modulator, with the exception that the input data are not split by serial to parallel converter, but the same input data is fed to the I and Q channels. The I, Q baseband signals and input data will all have the same information and bit rate. Such configuration creates a quadrature component and takes advantage of the FQPSK processor described earlier. The signal states of FBPSK avoid passing through the zero amplitude during the 0–180° state transition and vice versa, due to the added quadrature component. The FBPSK spectral properties have been evaluated both theoretically and experimentally, and are shown to exhibit low power-spectral properties, see Fig. 20. It can deliver 1-W output power compared to 150 mW for linear BPSK [25]. FBPSK constellation points and eye diagram are shown in Fig. 21; note the effect of the quadrature components and the FQPSK processor on the signal transition, where the zero signal amplitude is avoided during the transition from -1–$+1$. FBPSK modulated signals can be demodulated both coherently and noncoherently.

FBPSK modems have been evaluated experimentally using 2.4-GHz RFIC's, and have shown over 6-dB gain in output power over BPSK modulated systems [27]. In BPSK modulated systems, the output power has to be backed off by 6 dB in order to meet the 30-dB point of the FCC mask for WLAN as standardized by IEEE 802.11 committee. A BPSK receiver can recover FBPSK modulated signals with maximum degradation of 0.5 dB in an AWGN channel.

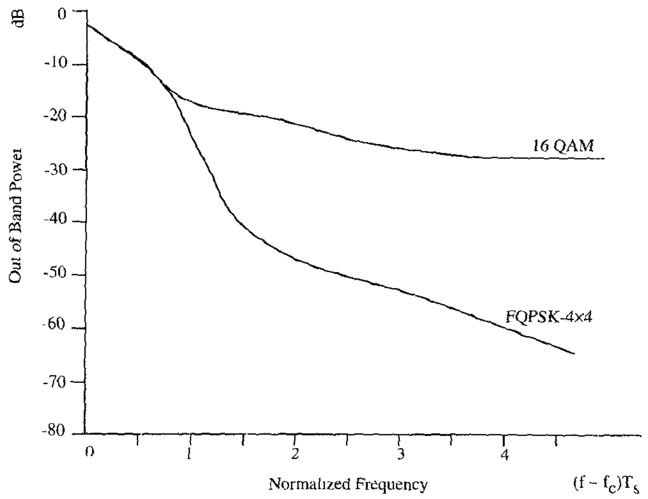


Fig. 23. Integrated out-of-band power of 16-QAM and FQPSK4 × 4 radio NLA Channel.

4) FQPSK4 × 4: To achieve a higher spectral efficiency in b/s/Hz, while maintaining power efficiency, a new, nonlinearly amplified modulation named FQPSK4 × 4 is presented. FQPSK4 × 4 is based on the FQPSK processor. A block diagram of this modulator is shown in Fig. 22, where the I and Q channels are amplified with two different NLA amplifiers (which have an output power differential of 6 dB) before being added for transmission. The power spectral density in the nonlinearly amplified channel is shown in Fig. 23. This modulation technique can achieve a spectral efficiency of 3 b/s/Hz in nonlinear amplified systems [22], [28], and can be demodulated using a conventional 16 QAM demodulator.

G. System Performance

The Bit Error Rate performance (BER) of the above described modulation techniques has been evaluated in

TABLE III
BER = $f(C/I)$ IN RAYLEIGH FADING AND BER = $f(E_b/N_o)k$ IN AWGN (STATIONARY) FOR GFSK, FQPSK, AND 4-FM CONSTANT ENVELOPE NLA SYSTEMS. THE $\pi/4$-DQPSK BER PERFORMANCE IS SIMILAR TO THAT OF FQPSK, HOWEVER, IT REQUIRES LINEAR AMPLIFIERS

	GFSK	FQPSK	QPSK	4-FM
Bit rate in 1MHz (-20dB)	1.0MB/s	1.6Mb/s	1 6Mb/s	2Mb/s
RF power@ 2.4GHz (max)	1 Watt (NLA)	1 Watt (NLA)	150 mW (Linear)	1 Watt (NLA)
Required C/I for BER = 10^{-2} Rayleigh	20dB	16dB	16dB	23dB
Increase in peak radiation	0dB	0dB	5dB to 10dB	0dB
Capacity (relative to GFSK)	100	300	300	50

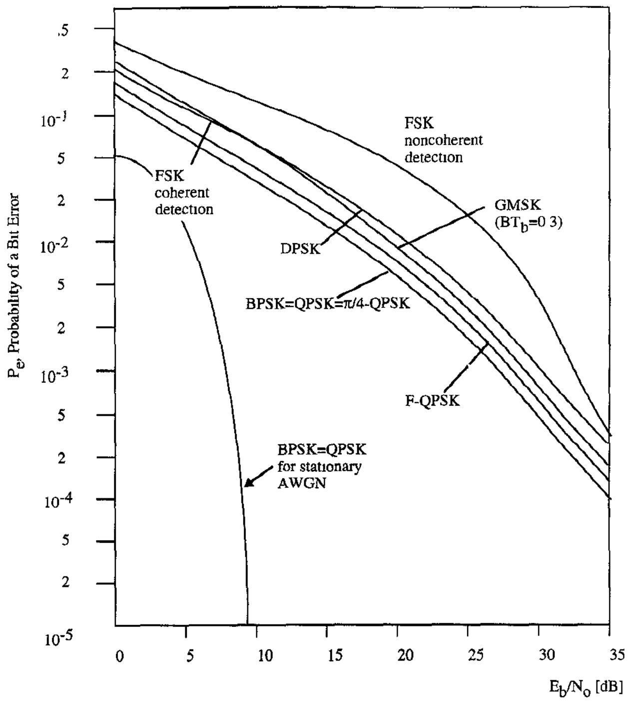


Fig. 24. $P_e = f(E_b/N_o)$ performance of linearly amplified coherent BPSK, QPSK, π/4-QPSK, FSK and of noncoherent DPSK and FSK in a slow Rayleigh faded channel. Nonlinearly amplified coherent GMSK and F-QPSK is also illustrated.

linear, nonlinear, Additive White Gaussian Noise (AWGN), and Rayleigh-fading channels. Coherent and noncoherent modulation methods are described and compared.

1) Bit Error Rate Performance: Linearly amplified systems require linear amplifiers, up/down converters, and linear Automatic Gain Control (AGC) circuits. The theoretical $P_e = f(E_b/N_0)$ performance of these systems is the best, however, linear amplification is not as power efficient (RF power/battery power ratio) and/or is considerably more complex and expensive than nonlinear amplifier and nonlinear AGC circuit implementations. Conventional QPSK, DQPSK, and π/4-QPSK systems require linear amplifiers, because in a nonlinearly amplified system their spectrum spread into adjacent channels.

Murota and Hirade have evaluated the performance of nonlinearly amplified MSK and GMSK systems in a stationary AWGN environment [17], [18]. These computer simulated results indicate that an "optimal" $BT_b = 0.25$ filtered

TABLE IV
ANALOG CELLULAR TELEPHONE STANDARDS [1]

STANDARD	AMPS Advanced Mobile Phone Service	TACS Total Access Communication System	NMT Nordic Mobile Telephone
Mobile Frequency Range (MHz)	Rx: 869-894 TX. 824-849	ETACS Rx. 916-949 Tx. 871-904 NTACS. Rx: 860-870 Tx: 915-925	NMT-450 Rx: 463-468 Tx: 453-458 NMT-900 Rx· 935-960 Tx: 890-915
Multiple Access Method	FDMA	FDMA	FDMA
Duplex Method	FDD	FDD	FDD
Number of Channels	832	ETACS: 1000 NTACS. 400	NMT - 450. 200 NMT - 900: 1999
Channel Spacing	300 kHz	ETACSL 25 kHz NTACS. 12.5 kHz	NMT - 450: 25kHz NMT - 900· 12 5 kHz
Modulation	FM	FM	FM
Bit Rate	n/a	n/a	n/a

GMSK coherent system is degraded approximately 1.6 dB, at BER $= 10^{-4}$, in comparison to ideal linearly amplified QPSK. The performance of nonlinearly amplified (as the simplest nonlinear amplifier a "hard-limited" amplifier has been used) FQPSK and of GMSK systems, and linearly amplified QPSK are shown in Fig. 16. In this figure, we note that at $P_e = 10^{-4}$, FQPSK and GMSK systems have very similar degradation, while at $P_e = 10^{-2}$, FQPSK is about 0.7 dB better than $BT_b = 0.3$-filtered GMSK [23]–[25]. This 0.7 dB advantage is significant in Rayleigh faded channels, where the bit error rate threshold is 10^{-2} for mobile cellular applications.

The probability of bit error P_e performance as a function of E_b/N_0, or $P_e = f(E_b/N_0)$ of ideal BPSK, FSK, QPSK, and of π/4-QPSK coherent systems, as well as of noncoherent DPSK and noncoherent FSK systems, is illustrated in Fig. 24. The performances of nonlinearly amplified FQPSK and GMSK systems (with $BT_b = 0.3$) coherent demodulation are illustrated by dotted lines. We note that at high E_b/N_o coherent PSK is 3 dB better than DPSK in a slow Rayleigh fading channel.

Table II summarizes the performances of GFSK, FQPSK, QPSK, and 4-level FM in AWGN and Rayleigh-fading systems. It shows that FQPSK increases capacity by 300% over GFSK and has over 5 dB output power advantage over QPSK. Spectral efficiency comparison of linearly and nonlinearly amplified QPSK, GMSK, GFSK, and FQPSK systems are presented in Table III.

TABLE V
DIGITAL CELLULAR TELEPHONE STANDARDS [1]

STANDARD	ADC: IS-54 North American Digital Cellular	IS-95 North American Digital Cellular	GSM Global System for Mobile Communications	JDC: PDC Personal Digital Cellular Japanese Digital Cellular
Mobile Frequency Range (MHz)	Rx: 869-894 Tx: 824-849	Rx: 869-894 Tx: 824-849	Rx: 935-960- Tx. 890-915	Rx: 940-956 Tx: 810-826 Rx: 1477-1501 Tx: 1429-1453
Multiple Access Method	TDMA/FDM	CDMA/FDM	TDMA/FDM	TDMA/FDM
Duplex Method	FDD	FDD	FDD	FDD
Number of Channels	832 (3 users/channel)	10 (118 users/channel)	124 (8 users/channel)	1600 (3 users/channel)
Channel Spacing	30 kHz	1250 kHz	200 kHz	25kHz
Modulation	π/4 DQPSK	BPSK/OQPSK	GMSK (0.3 Gaussian Filter)	π/4 DQPSK
Bit Rate	48.6 kb/s	1.2288 Mb/s	270.833 kb/s	42 kb/s

2) Coherent Demodulation Advantages over Noncoherent Systems: The performance and implementation complexity of coherent and noncoherent QPSK, F-QPSK and GMSK modulation/demodulation techniques in a complex mobile system environment, including large Doppler shift, delay spread, and low C/I, are compared. For large $f_d T_b$ products, where f_d is the Doppler shift and T_b is the bit duration, noncoherent (discriminator detector or differential demodulation) systems have a lower BER floor than their coherent counterparts. For significant delay spreads, e.g., $\tau_{\rm rms} > 0.4T_b$ and low C/I coherent systems outperform noncoherent systems. However, the synchronization time of coherent systems is longer than that of noncoherent systems [1], [25], [30].

Coherent systems have a simpler overall architecture (IF filter implementation-cost versus carrier recovery) and are more robust in an RF frequency drift environment [1], [10]. Additionally, the prediction tools, computer simulations, and analysis of coherent systems is simpler. In conclusion, we note that coherently demodulated radio/modem receivers have significant performance advantages compared to noncoherent systems.

III. MICROWAVE COMPONENTS

Chip makers are scrambling to serve the potential huge wireless market with Integrated Circuits (IC) that meet the variety of wireless communications standards being established around the world. Among these ICs are RF circuits for both analog and digital systems. All wireless systems, whether analog or digital, require RF/IF transceiver front-ends, including mixers, and low noise amplifiers (LNA). At this stage of development, technology is more resistant to higher levels of integration. There is also the competition between silicon and Gallium Arsenide when designing such RF components [33], [34]. In this section, a description of different RF and Microwave development technologies with a brief comparison of their advantages and disadvantages is presented. The impact of a chosen application or standard on the selection and design of RF components is briefly described. Finally, an evaluation of 2.4 GHz RFIC modems using FQPSK, FBPSK, DQPSK, BPSK, and PI/4 QPSK is presented.

A RF and Microwave Technologies

In Microwave and RF technologies, there are currently two leading technologies which have been targeting wireless applications, Silicon (Si),and Gallium Arsenide (GaAs) [33], [34]. The choice between silicon bipolar and Gallium Arsenide is not always obvious. Cost is the dominant parameter, but other factors such as noise, linearity and efficiency of amplifiers, as well as size, are important factors in the design selection. GaAs wafer material cost as much as 10 times that of a silicon, but requires fewer mask steps, and both cost the same in terms of testing and packaging. The real advantage of GaAs is in power amplifiers, GaAs amplifiers exhibit lower DC power drain compared to Silicon, and achieve higher power efficiency particularly at higher frequencies [35]. This is attractive for cellular and cordless telephone, and WLAN applications, where battery life and talk time are extremely important. In the ISM band of 2.4 GHz and above, GaAs is the preferred choice. Silicon, on the other hand is cheaper, and its major target is the 900 MHz band applications. It provides strong competition to GaAs where there is need for a low cost product with acceptable noise figure such as cordless telephone, compared to critical applications like base stations and satellite applications, where GaAs is preferred [36].

Other technologies have been developed to target different wireless parameters such as power efficiency, low power dissipation, low cost, small size, and low noise. Some of these technologies include the Siemens B6HF, by Siemens which is based on silicon bipolar technology; Siemens claims where they claimed it to be superior to both silicon and GaAs in terms of frequency operating range and cost [37]. Another technology developed by IBM and Analog Devices is based on silicon germanium (SiGe) and is tailored for low voltage, low power consumption RF and mixed signal applications with operating frequencies up to 5700–5800 MHz, which is the future ISM band [38]. To offset the high cost of GaAs, Motorola has developed a new technology named MAFET IC process [39]. This planar MESFET process borrows heavily from high-volume silicon processing and avoids the typical low-yield processes found in recesses-gate MESFET manufacturing. The 0.8-μm gate-length process, which includes the usual array of monolithic components, such as radial inductors, transmission lines, capacitors, and resistors, yields both enhancement and depletion of FETs for digital and RF applications.

B. RFIC, Microwave Components, and Applications

Standards can have a major impact on the direction of technologies and future applications. In wireless communications, many standards have been established to help achieve intentional compatibility among manufacturers, to accommodate frequency allocations, technologies, and help speed up the growth of new applications [1]. Some of the major standards in wireless communications including DECT, GSM, IS-54, and IS-95 are shown in Appendix A [40]–[44]. They specify many parameters which have impacted the RF and Microwave field development. Some of these parameters are: modulation type, frequency allocation, frequency operating range, interference specifications, output power, and error performance.

The above standard requirements coupled with need for products that are portable, light weight, and robustness have pushed RF technologies toward the development of integrated RF chip set. An integrated RF chip set design includes the sensitivity, dynamic range, spurious characteristics, and spectral shaping of the transmitted and received signal. A poorly designed chip can adversely effect the performance of the entire system performance. RF engineers should have an understanding of the system architecture in order to have an optimum system performance. Typical parameters are

1) Modulation scheme
2) Duplex method
3) Desired spectral shape and spurious dynamic range
4) receiver sensitivity for specific BER
5) Interface conditions and power management requirements.

The modulation technique chosen has an impact on the amplifiers chosen. Constant envelope modulations such as GMSK, FM, FQPSK, FBPSK allow the transmitter to operate in saturation, leading to higher output power and more robust system without spectral regeneration. Linear modulation such as DQPSK, PI/4-DQPSK, and BPSK have to operate in linear channel, and output power has to be backed off, resulting in less power efficient systems [1].

The transmission method such as half duplex or full duplex have a role in the system performance. In half duplex systems such as Digital European Cordless Telephone, or Personal Handy Phone (PHP), The Transmit /Receive Switch (T/R) plays a fundamental role in the system performance providing a low insertion transmission loss, and low distortion transmission path.

The upconverting mixer requires good Local Oscillator (LO) suppression to allow easy filtering with low IF frequencies. LO suppression as high as 20 dBc at the mixer output can be achieved for linear modulation schemes, and the number can reach 30 dBc for constant envelope schemes where the mixer can operate close to compression.

In Frequency Shift Keying (FSK) applications such as the frequency spectrum spread spectrum (FH-SS). The VCO is very critical; its frequency shift must be kept to a minimum relative to the frequency deviation of the FSK signal. Phase noise is also critical in frequency agile systems [1], [11].

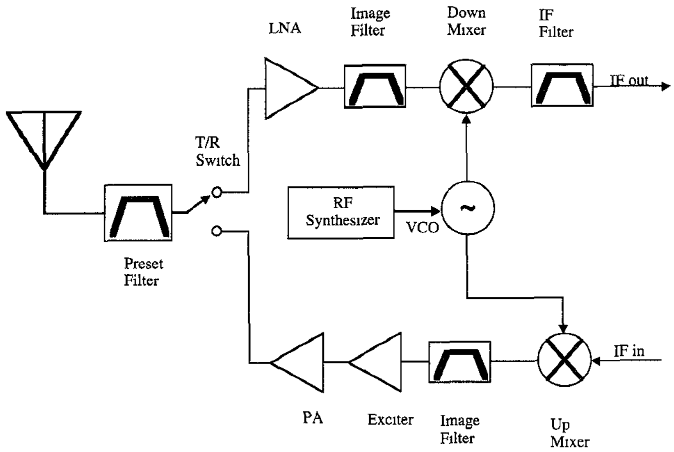


Fig. 25. Block diagram of a simplified 2.4-GHz radio [1].

The effect of battery life time, which impacts talk time is significant in portable and wireless communications systems. Systems which exhibit constant envelope properties can be nonlinearly amplified without output backoff. They can achieve up to 10 dB in output power advantage over linear modulated systems [1]. This translates into reduced power consumption, meaning longer battery life. The disadvantage in output power backoff and an additional 4–6 dB for design and components tolerance, gives constant envelope modulation around 6–10 dB in power advantage over linear modulated systems. In π/4-DQPSK, the peak to average power is about 4 dB compared to 0 for constant envelope schemes such as GMSK, FQPSK, FBPSK. This difference in average to peak power has an effect on the radiation as explained here: To achieve the same average output power, linear modulation schemes such as QPSK, and π/4 DQPSK must transmit more peak power than constant envelope systems; higher peak power means more electromagnetic radiation emission which may contribute to health hazards like cancer and Alzheimer diseases. FQPSK and FBPSK have been shown to demonstrate over 300% reduction in radiation emission compared to π/4-DQPSK, BPSK, and QPSK based system [31].

C. Experimental Evaluation at 2.4 GHz

A block diagram of a 2.4-GHz wireless LAN system is shown in Fig. 25. This system was evaluated using several Motorolas 2.4-GaAs RF integrated circuits, including the MRFIC 1801 antenna switch, MRFIC 1803 upconverter, MRFIC 1804 combined low-noise amplifier and down mixer, MRFIC 2401 downconverter, and MRFIC 2403 power amplifier. These RFIC's feature low-current operation at low voltages.

Using the 2.4-GHz configuration, the power spectral density of FQPSK, DQPSK, and FQPSK-KF has been evaluated in both linear and nonlinear amplified channels for WLAN applications [1], [27]. To meet the FCC mask requirement at −30-dB point for DS-SS WLAN applications, the power amplifier using DQPSK modulation has to operate in a linear channel, resulting in output power of 17.5 dBm. This compares to an output power of 24 dBm for FQPSK, a gain of almost

TABLE VI
DIGITAL CORDLESS TELEPHONE/PCN [1]

	CT2/CT2+ Cordless telephone 2	DECT Digital European Cordless Telephone	PHP Personal Hand Phone	DCS1800
Mobil frequency range	CT2: 864/868 CT2+: 930:931 940:941	1880-1990	1875-1909	Tx: 1710-1780 Rx: 1805-1880
Mobil access method	TDMA/FDM	TDMA/FDM	TDMA/FDM	TDMA/FDM
Duplex Method	TDD	TDD	TDD	FDD
# of channels	40	10 12 users/channel	300 4 users/channel	750 16 users/channel
channel spacing	100 KHz	1.728 MHz	300 KHz	200 KHz
Modulation	GFSK $BT_b = 0.5$ Gaussian filter	GFSK $BT_b = 0.5$ Gaussian filter	π/4-DQPSK	GMSK $BT_b = 0.3$ Gaussian Filter
Bit Rate	72 kb/s	1.152 Mb/s	384 kb/s	270.833 kb/s

TABLE VII
IEEE 802.11 SUMMARY OF PHYSICAL LAYER CHARACTERISTICS. 2.4 GHz DIRECT SEQUENCE SPREAD SPECTRUM (DSSS) [1]

Frequency	US 2.402-2.482 GHz Japan 2.471-2.497 GHz Europe 2.402-2.482 GHz
Spreading Sequence	11 chip Barker sequence +1,−1,+1,+1,−1,+1,+1,+1,−1,−1,−1
Modulation	1 Mb/s DBPSK 2 Mb/s DQPSK
Transmit Channel Spectrum Mask	−30 dBc @ f=±11 MHz −50 dBc @ f=±22 MHz
Preamble Length	160 bits
TX/RX Turnaround Time	25 μsec
RX/TX Switch/Ramp Time	10 μsec
Receiver Input Sensitivity	−80 dBm @ 10^{-5} BER
Bit Error Rate	10^{-5} @ Eb/No=17 dB

6.5 dB over DQPSK. Similar measurements were performed on BPSK and FBPSK, with BPSK showing an output power of 17.4 dBm compared to 24 dBm for FQPSK. This severe 6.5-dB output backoff is attributed to the envelope fluctuations of conventional BPSK and DQPSK systems. Such an increase in output power efficiency results in systems that reduce power consumption and extend battery life by more than 300%.

Additional measurements using different RFIC's components including Teledynes TAE-1010A 2.4 GHz power amplifier were performed. Using a TAE-1010A which is based on GaAs MMIC technology and has an output power of 27 dBm, it was shown that FQPSK has over 6-dB output power advantage compared to linear amplified DQPSK [47]. Similar tests were performed using Minicircuits VNA-25 RF amplifiers on both FBPSK, FQPSK, BPSK, DQPSK, and $\pi/4$-DQPSK, and similar power gains were reported also [1].

FQPSK processors have been implemented using Very Large Scale Integrated (VLSI) circuits, and Field Programmable Gate Array (FPGA) [1], [46], [48]. They consume less than 35 mW of active power and less than 150 μWatt of standby power and are supported by leading manufacturers such as Siemens of Germany [49] and NTT of Japan [19], [21].

IV. CONCLUSION

Some of the latest technologies in both digital modulation techniques and microwave development have been described. A new family of constant envelope nonlinearly amplified modulations named FQPSK have been presented, and they have been shown to have over 6–10 dB advantage in output power compared to linear modulated systems. A study of the leading modulation in Europe (GMSK) has shown the crosscorrelation between the in-phase and quadrature channels. The impact of modulation technique and standards on the development and design of RFIC's have been presented. Experimental evaluation using 2.4-GHz modems for WLAN applications have shown the advantages of FBPSK, and FQPSK over linear modulations in output power efficiency, battery life time, and reduction of output power.

APPENDIX A

In this Appendix, a selection of major worldwide standards are listed. It includes Analog Cellular standards such as AMPS in the US. Digital cellular standards including the IS-54 and IS-95 in the US, GSM in Europe, and JDC in Japan. The Digital Cordless Telephones, and Wireless LAN standards were adopted recently.

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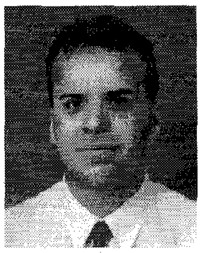
Kamilo Feher (M'68–S'73–M'73–SM'77–F'89) Professor at University of California, Davis, author of six books and of more than 300 R&D papers, is inventor of filter, signal processing and modulation patents licensed for wireless, cable, satellite and other communication and electronic systems. His patents are used for ROM, digital synthesized FPGA, ASIC Gaussian (e.g., GMSK and GFSK), raised cosine $\pi/4$-QPSK as well as other filter implementations, and in internationally standardized GSM system implementations of GMSK Quadrature correlated signal processors/modulators and GFSK systems. Feher's FQPSK, FBPSK, and "F-Modulation" family of patented nonlinearly amplified, combined modulation/RF and infrared wireless systems increased by 300% the throughput of WLAN standardized and other GFSK- and GMSK-compatible systems and have been proven to triple the RF output power and increase dramatically the battery power efficiency of QPSK type of systems. Through Dr. Feher Associates and as President of Digcom Inc. and of the International FQPSK Consortium, he is active in professional training courses, technology transfer, consulting and licensing of power and spectrally efficient wireless, cable and other filter/modulation RF products. At UC Davis he is directing one of the most active and best-equipped experimental digital and wireless communications university-based laboratories in the U.S. in the fields of CDMA, TDMA, FDMA modulation, and IF-RF design. Dr. Feher is Book Series Editor of *Feher/Prentice-Hall: Digital and Wireless Communications Series*. His most recent book *Wireless Digital Communications: Modulation and Spread Spectrum Applications* was published by Prentice-Hall, Englewood Cliffs, NJ 07632, 1995.



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In 1994, he joined American Microsystems as an Applications Engineer in the Telecom Group. In 1984, he joined Avantek Digital Radio Modem Group as a Design Engineer for microwave, and between 1987 and 1992 he was with Northern Telecom and Rolm Systems as a Design Engineer. In that same year, he joined the Digital Communications Research Laboratory where he worked on wireless personal communications systems. Recently, he rejoined Northern Telecom where he is working in the design group. His current research interests are in modulation, diversity, compatibility, and nonlinear amplified systems.