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Design and analysis of synchronous dynamic frequency hopping code division multiple access communication system

Mohammad M.N. Hamarsheh [a,*], Mohamad K. Abdullah [a], S. Khatun [a], Hossam M.H. Shalaby [b]

[a] *Computer and Communication Systems Engineering, University Putra Malaysia, Malaysia*
[b] *Department of Electrical Engineering, University of Alexandria, Egypt*

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Abstract

An optical dynamic frequency hopping code division multiple access communication system is proposed. In this system, an electrically controlled tunable optical filter (TOF) is used to encode the modulated broadband light source. The code depends on the function set to the controller. Two-dimensional code, named functional code, is also proposed based of shifted sine function. The function defines the dynamic coding pattern of the central wavelength of the transmitted narrowband optical signal. Thus, the system will allow for an easy reconfiguration of the transmitter without the need for sophisticated encoder. At the receiver, a synchronized TOF with the same function is used as a decoder. The system is modeled and analyzed taking into account the multiple access interference, phase induced intensity noise, and thermal noise. The performance of this system is shown to be better compared with a fast frequency hopping system and a spectral amplitude coding system that uses either a Hadamard code, a modified quadratic congruence code (MQC), or a modified frequency hopping code (MFH).

Keywords: Multiple access interference; Optical code division multiple access (CDMA)

1. Introduction

Many optical CDMA communication schemes have been proposed in the last two decades. The most attractive incoherent schemes are direct sequence (DS), spectral amplitude coding (SAC), and fast frequency hopping (FFH) optical CDMA systems. DS optical CDMA system encodes the incoherent pulses in time domain and recovers the data at the receiver using taped delay lines. The performance of this system is poor because of the correlation properties of the unipolar codes used which contributes to a high level of multiple access interference (MAI) [1]. SAC scheme is a more recent technique in optical CDMA systems where the spectrum of a broadband source is amplitude-encoded. In these systems, MAI can be canceled by balanced detection and code sequences with fixed in-phase cross correlation [2]. However, its performance is still limited by phase induced intensity noise (PIIN) [3,4]. FFH system was proposed in the late 1990s and it utilizes both time and frequency domains for encoding the optical signal [5]. Frequency separation between successive chip pulses is required in FFH-CDMA system to reduce the side lobe effects of the gratings. This limits the maximum number of users in the system. Furthermore, the spatial distance between the gratings and the number of gratings limits the users data bit rate in the system. Moreover, in all the above systems, they are either nonreconfigurable, or they needs complicated reconfigurable encoders [1,5,6].

In this paper we propose an easily reconfigurable dynamic frequency hopping optical CDMA (DFH-OCDMA) system. The encoder varies the central frequency of a pulse of optical signal according to the functional code set to the controller. The system can recover the encoded data by matched decoders at the receiver. In DFH-OCDMA, the TOF should be able to follow the functional code given as an electrical signal by the controller during one bit interval. The small data bit interval of the high data bit rate system requires fast TOF or special code with tuning range suitable with the speed of the TOF. How-

* Corresponding author. Current affiliation: Faculty of Information Science and Technology, Multimedia University, Malaka, Malaysia. Fax: +606 2318840.

E-mail address: hamarsheh@photronixm.com (M.M.N. Hamarsheh).

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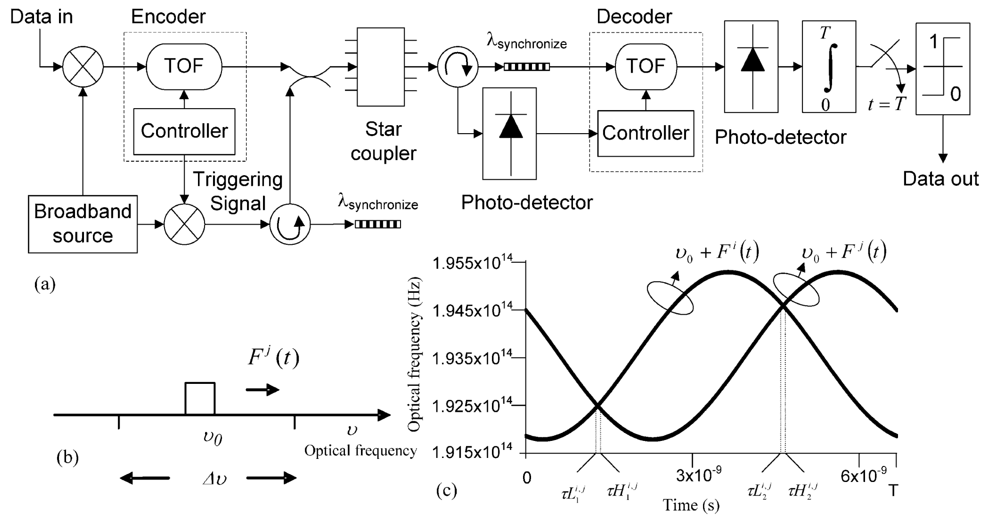

Fig. 1. (a) Block diagram of DFH-OCDMA system. (b) Optical spectrum of a signal from one of the users. (c) Power spectral density for two users as a function of time and frequency.

ever, tunable optical filters which can scan 10s of nanometers within few nanoseconds have been reported [7]. Thus, the encoder and decoder can be easily and quickly reconfigured to any of the functional codes. The implementation of the system will be shown to lead to better performance of the network. It is shown here that the system performance is better than that of SAC and FFH systems recently proposed [3–5].

2. System configuration and description

The block diagram in Fig. 1a shows the DFH-OCDMA configuration. The broadband signal from the light source is OOK modulated with the binary data. If the data bit is "1," encoder j, $j \in \{1, 2, \ldots, K\}$, where K is the number of simultaneous users, filters the spectrum of the pulse at a central wavelength that varies with time according to a functional code $F^j(t)$, otherwise no power is transmitted. The encoder is a tunable optical filter controlled with an electrical signal that represents the functional code. Part of the bandwidth ($\lambda_{\text{synchronise}}$) is reserved for synchronization between the transmitter and receiver TOFs. The controller sends a triggering signal to a modulator to produce a broadband triggering signal. Narrow bandwidth triggering signal is reflected from the fiber Bragg gratings filter (FBG) and combined with the data signal using 2:2 coupler. Signals transmitted from all synchronized users will be combined using a star coupler before being received by all users. Synchronizing optical CDMA system relaxes some of the features of CDMA to gain other advantages. It has advantages over asynchronous CDMA in terms of increased number of users [8–10]. At the receiver, the composite signal is decoded by a matched tunable optical filter. Another FBG filter at the decoder is used to recover the triggering signal which will be used by the controller to recover the data encoded in the broadband optical signal. Then, the signal passes through a photo-detector, an integrator, and a threshold decision to recover the transmitted data.

The source spectra are assumed to be flat over the bandwidth of $\upsilon_0 \pm \Delta\upsilon/2$, with magnitude $P_r/\Delta\upsilon$, where υ_0 is the central optical frequency, $\Delta\upsilon$ is the system bandwidth, and P_r is the received effective average power from a single source. Any excess losses in the route of the signal and the receiver are assumed to be incorporated in P_r. Ideal masking at the tunable optical filter is also assumed, and each user is considered to have the same effective average power at each receiver.

Figure 1b shows the spectrum of jth user's transmitted signal when the data bit is "1." It is similar to the spectrum of an ideal filter with central frequency varying with time according to a functional code. The proposed functional codes family $F(t)$ is shifted sine functions family with the same frequency and different phase shifts. Figure 1c shows an example of the spectrum for two users at the input of the decoder during one bit period when both users are sending a bit of "1." At the receiver side, the TOFs of the decoders are synchronized in time with a phase shift related to the functional code for each one of them. The output of the decoder is therefore the original signal which has the same phase shift of the decoder with some interference at the points of intersection with other users.

3. Code construction

The main criterion in the functional codes construction is to minimize the number of intersecting points between any pair of functions since they increase the interfering power between users. The area of intersection between any two functions is related directly to the value of interfering power and it is also important parameter in the construction of the functional codes. In our proposal, we have considered the use of shifted sine code (SSC) functions to alter the optical central frequency (υ_0) and to code the transmitted signal. The code family is given by

$$F^j(t) = \frac{\Delta\upsilon}{2} \sin(2\pi f t - j\varphi), \tag{1}$$

where f is the frequency of the functional code and φ is the phase shift between different functions. Shifted sine functions are proposed for their simplicity and the possibility of achieving the large number of required codes by reducing the phase shift.

The TOF in DFH-OCDMA should be able to follow the functional code driving the filter. The required speed of the TOF and its controller is defined as the derivative of the code and given by

$$S^j(t) = \Delta\upsilon\pi f \sin(2\pi f t - j\varphi). \tag{2}$$

It is directly proportional to the frequency and amplitude of the functional code. Thus, other codes could be proposed to improve the system performance and relax the implementation of the system for high data bit rates.

Furthermore, the functional codes should start and stop at the same central wavelength during the data bit interval (T) for smooth modulation of the TOF and its controller. This also limits the frequency of the code to be an integer value of $(1/T)$. For these reasons we use the smallest frequency possible for the SSC which equals to the data bit rate. Phase shift between codes (φ) is related to the spacing between users and the code size. Smaller phase shift results in a larger family of codes, but it reduces the spacing between users in the spectrum. The phase shift of SSC functions is chosen to be $2\pi/169$, that results of 169 different codes which is the same as the cardinality of MQC family of codes with $p = 13$ [2]. However, more users can be added to the system by assigning them new family of codes. These codes should be chosen with the same considerations given above. For example, another shifted sine functions with different amplitude will have only two intersecting points with any other code in the network, small area of intersection, and smooth modulation could also be achieved. Thus, more users can be added to the system and the system capacity is not limited to a specific number of users. Furthermore, transparency to user's data rate for flexible bandwidth provisioning can be achieved by using different code frequency for each data bit rate where higher data bit rates requires higher frequency of the SSC code.

4. DFH-OCDMA performance analysis

In the analysis of bit error rate (BER) we consider the effect of MAI, PIIN, and the thermal noise. Other sources, like shot noise and receiver's dark current noise are neglected. Gaussian approximation is assumed for the distribution of the noise in the calculation of the BER.

The variance of a photocurrent detected from unpolarized thermal light source generated by spontaneous emission including the effect of MAI can be expressed as

$$\sigma_t^2 = (K-1)\sigma_{\mathrm{DAI}}^2 + I^2 B\tau_c + 4k_B T_n B/R_l, \tag{3}$$

where $(K-1)\sigma_{\mathrm{DAI}}^2$ is the variance of the MAI, σ_{DAI}^2 is the variance of the interference when two users access the network, I is the average photocurrent, B is the noise-equivalent electrical bandwidth of the receiver, τ_c is the coherence time, k_B is the Boltzmann constant, T_n is the absolute receiver noise temperature in Kelvin, and R_l is the receiver load resistor. The first term of this equation represent the MAI effect. The second term denotes the effect of PIIN, where, incoherent light sources are mixed at the input of the photodetector and cause intensity variations of the output current. Finally, the third term represents the effect of thermal noise.

The power spectral density $G(\upsilon, t)$ of the signal at the input of receiver m, $m \in \{1, 2, \dots, K\}$ is the sum of all active users transmitted signals,

$$G_m^e(\upsilon,t) = \frac{P_r}{\Delta\upsilon}\sum_{j=1}^{K} b^j \operatorname{rect}\left(\frac{\upsilon - \upsilon_0 - F^j(t)}{\mathrm{BW}}\right),$$
$$\operatorname{rect}\left(\frac{\upsilon-\upsilon_0}{\mathrm{BW}}\right) = u\left(\upsilon - \upsilon_0 + \frac{\mathrm{BW}}{2}\right) - u\left(\upsilon - \upsilon_0 - \frac{\mathrm{BW}}{2}\right), \tag{4}$$

where $u(\upsilon)$ is the unit step function, BW is the bandwidth of the TOFs, and b^j is the data bit value of user j.

The receiver applies a synchronized matched TOF in decoding the incoming signal to extract the desired users data bit stream. The decoder output is:

$$\begin{aligned} G_m^d(\upsilon,t) = {} & \frac{P_r}{\Delta\upsilon} b^m \operatorname{rect}\left(\frac{\upsilon-\upsilon_0-F^m(t)}{\mathrm{BW}}\right) \\ & + \left(\frac{P_r}{\Delta\upsilon}\sum_{j=1, j\neq m}^{K} b^j \operatorname{rect}\left(\frac{\upsilon-\upsilon_0-F^j(t)}{\mathrm{BW}}\right)\right) \\ & \times \operatorname{rect}\left(\frac{\upsilon-\upsilon_0-F^m(t)}{\mathrm{BW}}\right). \end{aligned} \tag{5}$$

Then, the photo-current is:

$$\begin{aligned} I_m(t) = {} & \Re\int_{\upsilon=0}^{\infty} G_m(\upsilon,t)\,d\upsilon = \Re\frac{P_r}{\Delta\upsilon} b^m \mathrm{BW} \\ & + \Re\frac{P_r}{\Delta\upsilon}\sum_{j=1, j\neq m}^{K} b^j \sum_{i=1}^{N_{m,j}} \left(\mathrm{BW} - \left|F^m(t) - F^j(t)\right|\right) \\ & \times \left(u\left(t - \tau L_i^{m,j}\right) - u\left(t - \tau H_i^{m,j}\right)\right), \end{aligned} \tag{6}$$

where $\Re = (\eta e)/(h\upsilon_0)$ is the responsivity of the photo-detector, here η is quantum efficiency, e is the electron's charge, h is the Planck constant, $N_{m,j}$ is the number of intersecting points between users m and j during one bit period, and $\tau L_i^{m,j}$, $\tau H_i^{m,j}$ are defined as the roots of the following equations, respectively (see Fig. 1c):

$$F^m(t) - F^j(t) - \mathrm{BW} = 0, \tag{7}$$

$$F^m(t) - F^j(t) + \mathrm{BW} = 0. \tag{8}$$

After the integrator and sampler, the optical photo-current is:

$$\begin{aligned} I_m = {} & \frac{1}{T}\int_{t=0}^{T} I_m(t)\,dt = \Re b^m \frac{P_r}{\Delta\upsilon}\mathrm{BW} \\ & + \Re\frac{P_r}{T\Delta\upsilon}\sum_{j=1, j\neq m}^{K} b^j \sum_{i=1}^{N_{m,j}} \Bigg(\mathrm{BW}\left(\tau H_i^{m,j} - \tau L_i^{m,j}\right) \\ & - \int_{L_i^{m,j}}^{\tau H_i^{m,j}} \left|F^j(t) - F^m(t)\right| dt\Bigg). \end{aligned} \tag{9}$$

The optical photo-current at the receiver of user m, $m \in \{1, 2, \dots, K\}$ after the integrator and sampler can be reformulated as

$$I_m = b^m I + \mathrm{MAI}(m), \tag{10}$$

where $I = \Re P_\mathrm{r}\mathrm{BW}/\Delta\upsilon$, and the multiple access interference at receiver m, $\mathrm{MAI}(m)$ is given by

$$\mathrm{MAI}(m) = \sum_{j=0, j\neq m}^{K} \mathrm{DAI}(m, j), \tag{11}$$

where

$$\mathrm{DAI}(m, j) = \Re \frac{P_\mathrm{r}}{T\Delta\upsilon} \sum_{i=1}^{N_{m,j}} \left(\mathrm{BW}\big(\tau H_i^{m,j} - \tau L_i^{m,j}\big) - \int_{\tau L_i^{m,j}}^{\tau H_i^{m,j}} \left|F^j(t) - F^m(t)\right| dt \right) \tag{12}$$

is the interference between users m and j.

In Eq. (10), the first term is the data bit of the desired user m, and the second term is the MAI noise.

Since our system is synchronized, users m and j will interfere at the same points in time relative to the beginning of the bit period, and the intersecting edges $\tau L_i^{m,j}$ and $\tau H_i^{m,j}$ are the same whenever users m and j are active. This results in a constant value of $\mathrm{DAI}(m, j)$ if users m and j are active, otherwise $\mathrm{DAI}(m, j)$ is zero. The random and simultaneous access of the users, i.e., random m and j, results in a random value of the DAI, thus DAI is a random variable with average and variance given in Eqs. (13) and (14), respectively:

$$\mu_\mathrm{DAI} = \frac{1}{K^2 - K} \sum_{m=1}^{K} \sum_{j=1, j\neq m}^{K} \mathrm{DAI}(m, j), \tag{13}$$

$$\sigma^2_\mathrm{DAI} = \frac{1}{K^2 - K} \sum_{m=1}^{K} \sum_{j=1, j\neq m}^{K} \big(\mathrm{DAI}(m, j) - \mu_\mathrm{DAI}\big)^2. \tag{14}$$

The variance of MAI can be approximated as $(k-1)\sigma^2_\mathrm{DAI}$ for k simultaneous active users.

The PIIN causes variations in the output current during interference of incoherent light sources at the input of photodetector. The variance of the PIIN is related to the coherence time of the source (τ_c), as shown in Eq. (3), which is given by

$$\tau_\mathrm{c}(t) = \int_{\upsilon=0}^{\infty} G_m^2(\upsilon, t)\, d\upsilon \Bigg/ \int_{\upsilon=0}^{\infty} G_m(\upsilon, t)\, d\upsilon^2. \tag{15}$$

Assuming no more than one pair of users interfering at a time, which is the case in our proposed functional code family, averaging the variance at the points of interference along the bit period, and averaging over all users, the PIIN variance equation can be given by

$$\bar{\sigma}^2_\mathrm{PIIN} = \frac{1}{K} \sum_{m=1}^{K} \frac{1}{T} \int_0^T B\Re^2 \sum_{j=1, j\neq m}^{K} \sum_{i=1}^{N_{m,j}} \left(\left(\frac{P_\mathrm{r}}{\Delta\upsilon} b_m + \frac{P_\mathrm{r}}{\Delta\upsilon} b_j \right)^2 \times \big(\mathrm{BW} - \left|F^m(t) - F^j(t)\right|\big) + \left(\frac{P_\mathrm{r}}{\Delta\upsilon} b \right)^2 \left|F^m(t) - F^j(t)\right| \right) \times \Big(u\big(t - \tau L_i^{m,j}\big) - u\big(t - \tau H_i^{m,j}\big)\Big)\, dt. \tag{16}$$

The variance of the PIIN for k users can be expressed as $\sigma^2_\mathrm{PIIN} = k\bar{\sigma}^2_\mathrm{PIIN}$. From (3), (14), and (16), the signal-to-noise ratio can be expressed as

$$\mathrm{SNR}(k) = I^2 \Big/ \big\{(k-1)\sigma^2_\mathrm{DAI} + \sigma^2_\mathrm{PIIN} + 4k_\mathrm{B} T_\mathrm{n} B/R_\mathrm{l}\big\}, \tag{17}$$

and using Gaussian approximation, the BER is given by

$$\mathrm{BER}(k) = (1/2)\,\mathrm{erfc}\Big(\sqrt{\mathrm{SNR}(k)/2}\Big). \tag{18}$$

5. Results and discussion

The BER for DFH-OCDMA using proposed sine functional code family and another two OCDMA systems, one is FFH and the other is SAC system using either Hadamard code, MQC code with $p = 13$ [2], or MFH code with $q = 16$ [3], are plotted in Fig. 2 for the sake of comparison. It shows the relation between the BER and the number of simultaneous active users when $P_\mathrm{r} = -10$ dB m.

In our calculations we take $\Delta\upsilon = 30$ nm, $\upsilon_0 = 1550$ nm, BR = 155 Mbps, and filter bandwidth BW = 0.165 nm which is equal to the chip width of SAC system using MQC with $p = 13$ and same optical bandwidth.

For an error rate of 10^{-11}, DFH-OCDMA can accommodate up to 80 users, whereas for other systems, the maximum simultaneous users are 32 for SAC system using Hadamard code, 52 for SAC system using MQC code, 58 for SAC system using MFH code, and 24 for FFH system. The BER of the DFH-OCDMA system is increasing at a slower rate than that of the other two systems, which indicates that there is a significant improvement in performance at large number of users. Indeed it is shown that the BER for DFH-OCDMA is better that any other system at any number of users of more than 50. However, for less than 50 active users, SAC system with MFH or MQC gives BER better than that of DFH-OCDMA system. It should be noted that for this range of users, the error rate is very low ($<10^{-14}$).

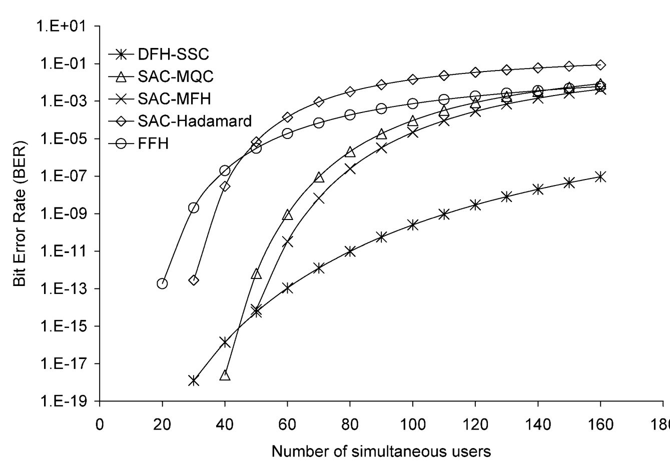

Fig. 2. Probability of error comparison between different optical CDMA systems.

6. Conclusion

We have proposed a novel low noise optical dynamic frequency hopping OCDMA communication system using a novel two-dimensional functional code. The encoder/decoder design is based on fast tunable optical filter. The filters are controlled dynamically and move one cycle during the data bit period. This encoder is easily reconfigured to any code by changing the electrical signal of the controller. The system is analyzed with a simple sine shifted functional code family taking into account the multiple access interference, the thermal noise, and the phase induced intensity noise. The system shows very small BER at large number of simultaneous active users compared with other systems like spectral amplitude coding and frequency hopping OCDMA systems. At 100 users, for example, the system BER is only 10^{-10}, while for all other systems it is higher than 10^{-5}. Although in the proposed system, the data transmission rate is limited by the tunable filter's tuning speed, other functional code families can be used whereby the requirement for tuning speed can be reduced so that the system can support higher bit rates.

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Mohammad M.N. Hamarsheh was born in Jenin, Palestine, in 1975. He received the B.Sc. degree in electrical engineering from An-najah National University, Nablus, Palestine, in 1999, and the M.S. degree in computer and communication engineering from University Putra Malaysia, Serdang, Malaysia, in 2002, where he is currently working toward the Ph.D. degree. He joined PhotronixM (R&D), Cyberjaya, in 2002, as a research engineer. He was involved in optical CDMA and fiber Bragg grating research and development. He is currently a lecturer at Multimedia University, Malaka, Malaysia. His research interests include optical code division multiple access systems, fiber Bragg gratings, dense wavelength division multiplexing, and polarization effect on optical fiber communication.

Mohamad Khazani Abdullah was born in Terenganu, Malaysia, in 1968. He obtained his B.Sc. and M.Sc. degrees in electrical engineering from University of Missouri at Rolla, USA, in 1990 and 1993, respectively, and Ph.D. degree from University Putra Malaysia, Serdang, Malaysia, in 1999. He was attached as Head, Photonics Devices Unit, Research and Development Division, Telecom Malaysia, from April 1996 to August 2000. He joined University Putra Malaysia, Serdang, Malaysia, as an Associated Professor in 1999. From July 2001, he has been the Head of Group 6, Malaysian National Top-Down Photonics project, Ministry of Science, Technology and Environment. He is now the Head of Computer and Communication Systems Engineering Department, Faculty of Engineering, University Putra Malaysia, Serdang, Malaysia. Currently he is the Chair of LEOS, Malaysian Chapter. His research interests include fiber optics devices, nonlinear optics, DWDM systems, and OCDMA systems. He is the recipient of Malaysian National Young Scientist Award for 2002, and has received various international invention shows including the Geneva and at INPEX, USA. He is currently holding eigth patents for various inventions.

Sabira Khatun received the B.Sc. (Hons.) and M.Sc. degrees in applied mathematics and Ph.D. degree in hydromagnetic stability in 1988, 1990, and 1994, respectively, all from the University of Rajshahi, Bangladesh. She received the second Ph.D. degree in communications and networking from University Putra Malaysia, Serdang, Malaysia, in 2003. She became a Lecturer at the Discipline of Computer Science and Engineering, University Khulna, Bangladesh, in 1991, and an Assistant Professor in 1994. She joined as Senior Lecturer at the Department of Computer and Communication Systems Engineering, University Putra Malaysia, Serdang, Malaysia, in 1998. She is an active researcher of Teman project and MyREN Research Community. She is a member of IEEE and her research interest spans broadband and wireless communications, and network management, including Software Defined Radio and IPv6.

Hossam M.H. Shalaby was born in Giza, Egypt, in 1961. He received the B.S. and M.Sc. degrees from the University of Alexandria, Egypt, in 1983 and 1986, respectively, and the Ph.D. degree from the University of Maryland, College Park, MD, in 1991, all in electrical engineering. In 1991, he joined the Department of Electrical Engineering, University of Alexandria, Egypt, as an Assistant Professor. He was promoted to the position of Associate Professor in 1996 and then to Professor in 2001 (current position). Since December 2000, he has been an Adjunct Professor with Department of Electrical and Information Engineering, Faculty of Sciences and Engineering, Laval University, Quebec, Canada. From March to April 1996, he was a Visiting Professor at the Electrical Engineering Department, Beirut Arab University, Lebanon. From September 1996 to January 1998, he was an Associate Professor with the Electrical and Computer Engineering Department, International Islamic University, Malaysia, and, from February 1998 to December 1998, he was with the School of Electrical and Electronic Engineering, Nanyang Technological University, Singapore, where he was a Senior Lecturer and, from January 1999 to February 2001, an Associate Professor. His research interests include optical communications, optical CDMA, spread-spectrum communications, and information theory. Dr. Shalaby received the SRC Fellowship from 1987 to 1991 (Systems Research Center, MD), the Shoman Prize for Young Arab Researchers in 2002 (The Abdul Hameed Shoman Foundation, Amman, Jordan), the State Award twice in 1995 and 2001 (Academy of Scientific Research and Technology, Egypt), and the University Award in 1996 (University of Alexandria, Egypt). He has served as a Student Branch Counselor (Alexandria University), IEEE Alexandria and North Delta Subsection, since 2002, and served as a Chairman of the Student Activities Committee of IEEE Alexandria Subsection from 1995 to 1996. He has also served as a technical referee for Proceedings of the Institution of Electrical Engineers, IEEE Transactions on Communications, IEEE Transactions on Information Theory, IEEE Journal on Selected Areas in Communications, and the Journal of Lightwave Technology. He is listed in the 14th edition of Marquis Who is Who in the World, 1997.